

PRACTICAL RADIO COMMUNICATION

Principles—Systems—Equipment—Operation

Including

Short-wave and Ultra-short-wave Radio

BY

ARTHUR R. NILSON

*Lieutenant (Technician) (Communications) U.S.N.R. (Retired);
Member of Institute of Radio Engineers*

AND

J. L. HORNUNG

*Formerly Radio Instructor, Guggenheim School of Aeronautics,
New York University; Member of Institute
of Radio Engineers*

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PREFACE

The modern radio communication center with its numerous transmitters and associated equipment is a far cry from the isolated radio station of a decade or so ago. The highly developed stations of the present day demand of their operating personnel a skill which is closely akin to radio engineering. Broadcasting and its consequent effect on all radio are largely responsible for this. Broadcasting gave a spurt to radio unequalled in the annals of any other scientific development. Apparatus was hurriedly built and the erection of stations rushed. The technical aspects of radio were in a constant upheaval, apparatus, methods, and even systems changing from month to month. But the initial confusion is now over; radio has found its stride, so to speak, and its technique is quite settled. Radio evolution in the future holds promise of being less spasmodic.

The situation thus pictured has been recognized by the government in the establishment of the Federal Communications Commission with a special division charged with the establishment of technical standards to which all licensed radio operators must conform. These standards are high and keep pace with scientific radio development. But there are other even higher standards up to which the technical radio operator-technician must measure. We refer to the standards set up by employers of technical radio personnel. These employers have found it good business to employ only thoroughly qualified men. It follows, therefore, that technical radio education must fill this demand.

It was to meet the expanded scope of technical radio requirements that this book was planned. We believe that our objective of providing within one volume most of the technical information required by the practical radio operator-technician has been achieved. The original manuscript would have made a book almost 50 per cent larger than the present volume, with a consequent increase in price. Hence, in order to keep the size of the book within reasonable limits, many sections were condensed. Wherever a diagram in the practical sections told the story, a word-for-word explanation was omitted. What the reader does not find in the text, therefore, he will usually learn from the diagram. As far as possible, items on the diagrams have been labeled. This has been done to aid the reader and instructor in referring to these specific items, even though a list of parts may not accompany the diagram.

The book is divided roughly into two parts: Principles and Practice. The first six chapters are given over to principles and the remaining nine

chapters to practice. High-grade technical skill is always founded on a sound basis of fundamental principles. It has been our purpose, therefore, to go into these principles extensively, delving deeply into alternating current, which plays an important part in modern radio. It is on the assumption that these principles have already been mastered that the practical sections of the book have been written. In fact, it is necessary that these principles be understood before the practical sections, beginning with Chap. VII, can be read intelligently.

Throughout the book, wherever current flow is treated, the electron viewpoint (negative to positive) has been adhered to. Indications are that the general trend in radio technical circles is in this direction, rather than in the direction of a blind acceptance of the conventional (positive to negative) direction law expounded in the majority of texts on electricity.

Particular attention has been given to broadcasting, and much of the material included on this subject appears here for the first time in a radio textbook. The practical description of the Western Electric broadcast transmitter is written on a basis of wide operating experience with this equipment. The treatment as here given is not a mere digest of instruction-book material; and the diagrams accompanying the text were especially prepared and simplified for our purpose.

Broadcast men will also find the chapters on Studio Acoustics and Apparatus, Control-room Equipment and Operation, and Antennas to be of great value for the ordinarily inaccessible information which they contain.

Police operators will find much of interest in the broadcast sections and; in addition, in the sections on ultra-short-wave equipments, which are now extensively used for police communications.

Aviation radio is treated from the radio-telegraphic, as well as from the radio-telephonic, angle. Modern aircraft transmitters, receivers, and direction-finding equipment together with the testing and maintenance of this apparatus are included.

Nowhere is radio more important than in its marine applications; and it is with this in mind that the sections dealing with marine radio have been written. In addition to the conventional medium-frequency equipment, high-frequency transmitters and receivers have been fully illustrated and explained. Low-power transmitters such as are used by the Coast Guard have also been included. Marine direction finders and the principles underlying all such equipment are completely covered in a special chapter.

The last three chapters of the book are given over to a thorough treatment of power-supply apparatus including rectifiers, generators, and batteries, and their associated equipment. By placing this material at the end of the book we have saved the reader the necessity of tediously

completing these power chapters before getting on to the more interesting study of radio circuits. We feel that this approach to the subject, although a new one, offers distinct pedagogical advantages.

Lack of space did not permit a discussion of the nearly obsolete spark and arc transmitters or of radio laws. Those readers who wish information on these old type transmitters will find them adequately described in the authors' "Practical Radio Telegraphy," the predecessor to this volume. Radio laws have been fully treated in the companion book to this volume, the authors' "Radio Operating Questions and Answers," and in inexpensive government publications.

In short this book covers the requirements for *all classes* of radio operator's license examinations, treats long, medium, short, and ultra-short-wave radio, includes *all classes* of radio stations, and, in general, is a complete text on the practical radio communication based on a theoretical introduction.

THE AUTHORS.

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CONTENTS

	PAGE
PREFACE	V
ACKNOWLEDGMENTS	ix
SYMBOLS AND ABBREVIATIONS	xix

CHAPTER I

DIRECT-CURRENT ELECTRICITY AND MAGNETISM	1
--	---

PART I. ELEMENTARY ELECTRICITY: 1. Electric Charges and Forces; 2. Protons and Electrons; 3. Potential; 4. Free Electrons in Solid Conductors; 5. Electric Current; 6. Electromotive Force; 7. Current Strength and Quantity; 8. Electrical Resistance and Conductivity of Materials; 9. Ohm's Law; 10. Simple Electric Circuits; 11. The Shunt Law; 12. Kirchhoff's Laws; 13. Conductance; 14. Energy, Force, Work, and Power; 15. Electrical Power; 16. Load Matching; 17. E. M. F. Produced by Chemical Action; 18. Directions of Current Flow; 19. Local Action; 20. The Voltage of a Cell; 21. The Capacity of a Cell; 22. Connection Methods; 23. The Resistances of a Cell; 24. Dry Cells; 25. The Telephone Receiver and Transmitter.

PART II. MAGNETISM AND ELECTROMAGNETISM: 26. Natural and Artificial Magnets; 27. Poles of a Magnet; 28. Magnetic Fields; 29. Magnetic Transparency and Leakage; 30. Theory of Magnetism; 31. Magnetic Variables; 32. Purpose of Pole Pieces; 33. Magnetostriction Effect; 34. Fields around Conductors; 35. Induced Currents; 36. Production of E.M.F. by Field Cutting Conductor.

PART III. THE MAGNETIC CIRCUIT: 37. The Magnetic Circuit; 38. Ohm's Law for Magnetic Circuits; 39. Magnetic-circuit Computations; 40. Hysteresis.

CHAPTER II

ALTERNATING-CURRENT ELECTRICITY	58
---	----

PART I. ALTERNATING-CURRENT GENERATION AND TRANSFORMATION: 41. The Alternator Principle; 42. Principles of the Transformer.

PART II. INDUCTANCE: 43. Induction; 44. Self-induction; 45. Mutual Induction; 46. Inductance Measurements; 47. Practical Inductance Formulas; 48. Mutual Inductance Measurements.

PART III. CAPACITY: 49. The Electrostatic Field and the Condenser; 50. Capacity of a Condenser; 51. Specific Inductive Capacity; 52. Energy in a Charged Condenser; 53. Capacity of Condensers in Parallel; 54. Capacity of Condensers in Series; 55. Capacity Measurements.

PART IV. ELEMENTS OF ALTERNATING-CURRENT ENGINEERING: 56. Construction of a Sine Wave; 57. Frequency and E. M. F. Formulas; 58. Effective Voltage and Current; 59. Effect of Resistance on Phase Relations; 60. Effect of Inductance on Phase Relations; 61. Effect of Capacity on

Phase Relations; **62.** Power Factor; **63.** Ohm's Law for Alternating-current Circuits; **64.** Phase Relations in Series-resonant Circuits; **65.** Phase Relations in Parallel-resonant Circuits; **66.** Parallel-circuit Analysis; **67.** Resonance; **68.** The *J* Operator; **69.** Power Relations in an Alternating-current Circuit; **70.** Kilovolt-amperes; **71.** Polyphase Alternating-current Circuits; **72.** Power in a Three-phase System.

CHAPTER III

INTRODUCTION TO VACUUM TUBES. 116

PART I. VACUUM-TUBE CHARACTERISTICS: **73.** The Filament; **74.** Effect of Plate Potential on Plate Current; **75.** Effect of Filament Current on Plate Current; **76.** The Control Grid; **77.** Vacuum-tube Characteristics; **78.** Amplification Factor; **79.** Plate Resistance; **80.** Mutual Conductance; **81.** The Alternating-current Operated Tube.

PART II. THE RADIO-FREQUENCY AMPLIFIER: **82.** General Amplifier Considerations; **83.** The Class A Amplifier; **84.** The Class B Amplifier; **85.** The Class C Amplifier; **86.** Balance Circuits; **87.** Neutralization of a Transmitting Circuit; **88.** Magnetic Shielding; **89.** Inductance and Capacity Changes with Shielding; **90.** Grid-biasing Methods; **91.** Radio-frequency Chokes.

PART III. VACUUM-TUBE OSCILLATORS: **92.** General Oscillator Considerations; **93.** Fundamental Oscillator; **94.** The Hartley Oscillator; **95.** The Colpitts Oscillator; **96.** The Push-pull Oscillator; **97.** Conventional Electron-coupled Oscillator; **98.** The Dynatron Oscillator.

CHAPTER IV

TRANSMITTING-CIRCUIT PRINCIPLES 155

PART I. CONVENTIONAL CIRCUITS: **99.** General Transmitting Considerations; **100.** Classification of the Radio Spectrum; **101.** The Oscillator as a Transmitter; **102.** The Oscillator-amplifier Transmitter; **103.** The Crystal-controlled Transmitter; **104.** The Practical Oscillator; **105.** The Mechanical Resonator; **106.** The Crystal Resonator; **107.** The Temperature-control Chamber; **108.** The Buffer Amplifier; **109.** The Second Amplifier; **110.** Harmonic Producers: Frequency Multipliers; **111.** Modulation; **112.** Modulation Percentage; **113.** Carrier and Side-band Relations; **114.** Modulation Methods; **115.** The Modulating Amplifier—Push-pull System; **116.** The Heising Modulating Systems; **117.** The Output Impedance of the Modulated Amplifier; **118.** The First- and Second-power Amplifiers; **119.** Power-amplifier Grid Excitation; **120.** Calculation of Load Impedance in Power-amplifier Systems; **121.** The Audio Monitor; **122.** The Frequency Monitor; **123.** Antenna Coupling Circuits; **124.** Meters; **125.** The Power Supply; **126.** Harmonics and Their Causes; **127.** The Suppression of Harmonics; **128.** Parasitic Oscillations.

PART II. ULTRA-HIGH-FREQUENCY CIRCUITS: **129.** Ultra-high-frequency Systems; **130.** Ultra-high-frequency Oscillators; **131.** Modulation of Ultra-high Frequencies.

CHAPTER V

RECEIVING-CIRCUIT PRINCIPLES. 207

PART I. THE DETECTOR AND AUDIO AMPLIFIER: **132.** General Receiving Considerations; **133.** The Screen-grid Tube; **134.** The Vacuum-tube Detector; **135.** Plate Detection; **136.** Grid Detection; **137.** Power Detection;

138. Grid Leak versus Power Detection; 139. The Vacuum-tube Amplifier; 140. Amplifier Distortion; 141. Power Amplification; 142. The Pentode Tube; 143. The Power Pentode Load; 144. Push-pull Amplification; 145. Grid and Plate Filtration; 146. The Diode Detector; 147. The Automatic Volume Control; 148. Noise-suppression Circuit.

PART II. THEORY AND APPLICATION OF TUNED CIRCUITS: 149. General Circuit Considerations; 150. The Theory of Coupled Circuits and Tuning; 151. Simple Analysis of the Tuned Circuit; 152. Filter Circuits; 153. Isolation and Band-pass Circuits; 154. Rejector or Trap Circuits; 155. Regenerative Receiver; 156. Superregenerative Receiver; 157. Superheterodyne or Double-detection Receiver; 158. The Single-signal Receiver.

CHAPTER VI

ANTENNAS AND WAVE PROPAGATION. 261

PART I. FUNDAMENTAL TYPES OF ANTENNAS: 159. The Half-wave Doublet; 160. The Quarter-wave Antenna.

PART II. ELEMENTS OF RADIATION: 161. The Radiation Field; 162. Sky and Ground Wave; 163. Field Intensity; 164. Field-strength Measurements; 165. Propagation of Short Waves; 166. Skip Distance; 167. Echo Signals and Multiple Reflections; 168. Propagation of Ultra-short Waves; 169. Polarization.

PART III. RADIATING SYSTEMS: 170. Short-wave Antenna Design; 171. Broadside Directional Antenna; 172. The Franklin Antenna; 173. The Diamond-type Antenna; 174. Feeding the Antenna; 175. Transmission Line as a Phase or Voltage Transformer; 176. Broadcast Transmission-line Coupling Circuits; 177. Application of the Antenna-coupling Unit; 178. Construction of Short-wave Antennas; 179. Ultra-high-frequency Antennas; 180. Broadcast-transmitting Antennas; 181. Ship Antennas; 182. Airplane Antennas; 183. Ground Connection and Counterpoise; 184. Receiving Antenna.

PART IV. ANTENNA MEASUREMENTS: 185. Measurements on Broadcasting Antennas; 186. Calculations on Antennas.

CHAPTER VII

STUDIO ACOUSTICS AND APPARATUS 309

PART I. STUDIO ACOUSTICS: 187. Distortion; 188. Reverberation; 189. Microphone Placement; 190. Sound; 191. Use of Logarithms; 192. The Decibel.

PART II. MICROPHONES: 193. Microphones—General Considerations; 194. Carbon Microphone; 195. Condenser Microphone; 196. Dynamic Microphone; 197. Velocity or Ribbon Microphone; 198. Crystal Microphone; 199. Lapel Microphone; 200. Parabolic Microphone Receiver-reflector.

PART III. ELECTRICAL TRANSCRIPTION APPARATUS: 201. Recordings; 202. Pick-up Heads; 203. Tone-arm Balance; 204. Pick-up Impedance and Output; 205. Scratch Filter; 206. Types of Microphones and Pick-ups Compared.

CHAPTER VIII

CONTROL-ROOM EQUIPMENT AND OPERATION 345

PART I. TRANSMISSION SET-UP: 207. Control-room Requirements; 208. Control-room Equipment.

PART II. MIXING EQUIPMENT: 209. General Requirements and Types of Mixers; **210.** Stray Fields and Noises; **211.** Insertion Loss; **212.** Special Mixing Problems; **213.** Adjunct (NEMO) and Pick-up Mixing.

PART III. STUDIO AMPLIFIERS: 214. Special Precautions against Disturbances; **215.** Harmonic and Amplitude Distortion; **216.** Shot Effect; **217.** Advantages of Low-gain Amplifiers; **218.** The Resistance-coupled Amplifier; **219.** The Impedance-coupled Amplifier; **220.** The Transformer-coupled Amplifier; **221.** The Main Gain Control; **222.** Bridging Circuits; **223.** Power Supplies; **224.** Biasing Voltage; **225.** Master Control Room.

PART IV. ALTERNATING-CURRENT OPERATED EQUIPMENT: 226. Advantages of Alternating-current System; **227.** Amplifier-tube Developments; **228.** Reducing Hum Level; **229.** Time-delay Relay; **230.** Voltage Regulation; **231.** Amplifiers in Alternating-current-operated Equipments.

PART V. PROGRAM CONTROL AND TRANSMISSION: 232. Monitor Amplifiers; **233.** Volume Indicators; **234.** Use of Volume Indicators; **235.** Equalizers and Equalizing; **236.** Frequency Runs; **237.** Remote-control Equipment; **238.** Attenuation Networks.

CHAPTER IX

BROADCAST TRANSMITTERS 391

239. The Broadcast Station; **240.** Definition and Rating of the Transmitter; **241.** The Operating Power; **242.** Frequency Control; **243.** Western Electric Broadcast Transmitter; **244.** Relay Operation; **245.** General Operation and Tuning; **246.** The Speech-input Equipment; **247.** The Frequency Monitor; **248.** Modulation Measurements; **249.** RCA Victor Broadcast Transmitters.

CHAPTER X

COMMUNICATION TRANSMITTERS. 421

PART I. MARINE TRANSMITTERS: 250. Marine Transmitters; **251.** Keying Systems; **252.** Production of *ICW*; **253.** Emergency Transmitter; **254.** High-frequency Transmitter; **255.** Medium-frequency Transmitter; **256.** Radiophone Attachment; **257.** Radio Telephone-telegraph Transmitter; **258.** A Complete Ship-radio Station; **259.** Transmitter Care and Maintenance; **260.** Transmitter Troubles; **261.** Coastal Stations.

PART II. LOW-POWER TELEPHONE AND TELEGRAPH TRANSMITTERS: 262. Radio-telephone Transmitter; **263.** Low-power Rectifier Unit; **264.** Converting a Radio-telephone Transmitter for *CW*; **265.** Aircraft Radio-telephone Transmitter; **266.** Aircraft Radio-telegraph Transmitter; **267.** Ultra-short-wave Transmitters.

CHAPTER XI

RADIO RECEIVERS 475

PART I. GENERAL-PURPOSE RECEIVER: 268. All-wave Receiver.

PART II. MARINE RADIO RECEIVERS: 269. Marine Receivers; **270.** The Coupling Adjustment; **271.** Audio-amplifier Response Characteristics; **272.** Audio-amplifier Troubles; **273.** Shortening Condenser; **274.** Power Supplies for Receivers; **275.** Protective Devices; **276.** Multiple Reception; **277.** High-frequency Receiver; **278.** Combination-frequencies Receiver; **279.** Medium-frequency Receiver; **280.** Long-wave Attachment; **281.** Common Receiver Troubles Summarized.

CONTENTS

XV
PAGE

PART III. AIRCRAFT AND POLICE RECEIVERS: 282. Aircraft Receivers: General Considerations; 283. Medium-high-frequency Aircraft-communications Receiver; 284. Radio-beacon and General-purpose Aircraft Receiver; 285. Antennas for Radio-range-beacon Reception; 286. Aircraft Shielding and Bonding; 287. Police Radio Receivers; 288. Shielding on Automobiles; 289. Ultra-short-wave Receivers.

PART IV. FREQUENCY MEASUREMENTS: 290. Heterodyne Frequency Meter.

CHAPTER XII

RADIO AIDS TO NAVIGATION. 523

PART I. PRINCIPLES OF RADIO DIRECTION FINDERS: 291. Fundamental Principles of the Loop Antenna; 292. Minimizing the Error; 293. Effect of Frequency on Deviation; 294. Determining Sense of Direction; 295. The Bellini-Tosi System; 296. The Adcock Aerial for Directional Reception.

PART II. THE ROTATING-LOOP TYPE RADIO DIRECTION FINDER: 297. Applications of the Radio Direction Finder; 298. Elements of Direction-finder Apparatus; 299. Grounds and Shielding; 300. Signal Indication; 301. Schematic Circuit Diagrams; 302. Calibration; 303. Operation; 304. Fixing the Position; 305. Maintenance; 306. Troubles and Remedies.

PART III. RADIO AIDS TO OVERLAND AIR NAVIGATION: 307. Radio-range Beacons; 308. Action of Loop Transmitting Antennas; 309. The T-L Antenna System; 310. Radio Beacon-signal Methods; 311. The Single-loop Direction Finder for Aircraft.

CHAPTER XIII

RECTIFIER UNITS. 575

PART I. THE POWER TRANSFORMER: 312. General Transformer Considerations; 313. Voltage-regulating Transformers; 314. Electrostatically Shielded Transformers; 315. Protective Devices.

PART II. THE RECTIFIER: 316. The Function of the Rectifier; 317. The High-vacuum Rectifier Tube; 318. The Mercury-vapor Rectifier Tube; 319. Inverse-peak Voltage; 320. Rectifier Circuits Classified; 321. Single-phase Rectifier Circuits; 322. Bridge-rectifier Circuits; 323. Voltage-doubler Rectifier; 324. Three-phase Rectifiers; 325. Mercury-arc Rectifiers; 326. Metal-clad Mercury-arc Rectifiers; 327. Junction Rectifiers.

PART III. THE RIPPLE FILTER: 328. Function of Ripple Filter; 329. The Action of the Condensers; 330. Paper Filter Condensers; 331. Electrolytic Filter Condensers; 332. The Action of the Choke Coil; 333. The First Choke Coil; 334. Equalizing and Bleeder Resistance; 335. Voltage Dividers.

PART IV. GENERAL RECTIFIER CONSIDERATIONS: 336. Ripple Frequency; 337. Losses in Rectifier Apparatus; 338. Typical Rectifier and Filter Systems; 339. Plate Fuses; 340. Rectifier Troubles.

CHAPTER XIV

DYNAMO-ELECTRIC MACHINERY AND METERS. 617

PART I. GENERATORS: 341. Types of Alternators; 342. Voltage Control on Alternators; 343. Direct-current Generators; 344. Commutation; 345. Commutator Sparking; 346. Commutating Poles; 347. Compensating-field Windings; 348. Double-commutator Generators; 349. The Drum Armature; 350. Armature Cores; 351. Field Excitation; 352. Brushes; 353. Voltage Control on Direct-current Generators; 354. Uses of Various Types of Generators; 355. Regulation.

PART II. MOTORS: **356.** Construction Principles of the D.-c. Motor; **357.** The Function of the Motor Commutator; **358.** Counter Electromotive Force: Torque; **359.** Types of Direct-current Motors; **360.** Alternating-current Motors.

PART III. MOTOR STARTING, CONTROL AND OPERATION: **361.** Hand-starting Boxes; **362.** Automatic Starters for Direct-current Motors; **363.** Starting Induction Motors; **364.** Manual Speed Control; **365.** The Centrifugal Speed Regulator; **366.** Operation of Motors.

PART IV. MOTOR GENERATORS AND POWER-CONTROL APPARATUS: **367.** Advantages and Disadvantages of Motor Generator; **368.** The Dynamotor; **369.** The Converter; **370.** Double-current Generators; **371.** General Maintenance; **372.** Brush Troubles; **373.** Bearings; **374.** Trouble Shooting; **375.** Protective and Control Devices; **376.** Generator-ripple Reduction; **377.** Generator-ripple Filters.

PART V. ELECTRICAL INDICATING INSTRUMENTS: **378.** Voltmeter for Continuous E.M.F.; **379.** Ammeter for Direct Current; **380.** Voltmeter for Alternating E.M.F.; **381.** Ammeters for Alternating Current; **382.** Thermal Ammeters; **383.** Wattmeters; **384.** Frequency Indicators; **385.** The Ohmmeter.

CHAPTER XV

STORAGE BATTERIES 671

386. General Battery Considerations.

PART I. THE LEAD ACID BATTERY: **387.** The Lead-acid Cell; **388.** Positive Plates; **389.** Negative Plates; **390.** The Active Material; **391.** The Electrolyte; **392.** Construction of Cells; **393.** Chemical Action in a Lead Cell; **394.** Specific Gravity and the Hydrometer; **395.** The Voltage of the Lead Cell; **396.** Discharge Rates and Limits; **397.** Charging and Charging Rates; **398.** Charging Requirements; **399.** Charging Methods; **400.** Effect of Temperature on Operation; **401.** Gas and Ventilation; **402.** Effect of Wrong Charging Polarity; **403.** Sulphation and Remedies; **404.** Care and Operation of Lead Cells; **405.** Troubles and Their Remedies.

PART II. THE NICKEL-IRON-ALKALINE BATTERY: **406.** The Nickel-iron-alkaline Cell; **407.** Care and Management; **408.** The Chemical Reaction in an Edison Cell; **409.** Charging; **410.** The State of Charge of an Edison Battery; **411.** Ventilation; **412.** Capacity; **413.** Summary of Data on Lead and Edison Cells.

PART III. STORAGE BATTERY SWITCHBOARDS AND CHARGERS: **414.** The Exide Switchboard; **415.** The Edison Battery Switchboard; **416.** Charging Panel for Receiver Batteries; **417.** Lifeboat Panel; **418.** Ampere-hour Meters; **419.** The Tungar Battery Charger.

APPENDIX A

USEFUL TECHNICAL INFORMATION CHARTS AND TABLES

THE NEW TUBE NUMBERING SYSTEM	713
TABLE X.—THE R.C.A. RADIOTRON-CUNNINGHAM RADIO-TUBE CHART	714
TUBE SYMBOLS AND BOTTOM VIEWS OF SOCKET CONNECTIONS	720
TABLE XI (PART 1).—CHARACTERISTICS OF TRANSMITTING RADIOTRONS, AMATEUR TYPES	721
TABLE XI (PART 2).—CHARACTERISTICS OF TRANSMITTING RADIOTRONS	724
TABLE XII.—CHARACTERISTICS OF RECTIFIER TUBES	727
TABLE XIII.—CHARACTERISTICS OF WESTERN ELECTRIC VACUUM TUBES	728

CONTENTS

xvii

PAGE

LEGENDS AND NOTES FOR TABLES XI, XII, AND XIII	730
TABLE XIV.—WAVE-LENGTH-FREQUENCY CONVERSION TABLE (VERY HIGH FREQUENCIES)	732
TABLE XV.—RELATION OF WAVE LENGTH (λ), FREQUENCY (f), $2\pi f = (\omega)$, AND $L \times C$	733
BIBLIOGRAPHY	737

APPENDIX B

USEFUL OPERATING INFORMATION

LIST OF ABBREVIATIONS TO BE USED IN RADIO COMMUNICATION	739
SCALE USED TO EXPRESS STRENGTH OF SIGNALS.	742
RADIO WEATHER TRANSMISSION CODE.	742
INTERNATIONAL MORSE CODE.	744
INDEX.	745
THREE-PLACE LOGARITHMIC TABLES.	756

SYMBOLS AND ABBREVIATIONS

Note: Where a single symbol is given more than one meaning in this list, the application in a particular instance is obvious from the context of the text.

- A = voltage amplification or gain.
- A = area or cross section; amperes.
- a.c. = alternating current; in tables and lists = A.-C.
- ACW = attenuated continuous wave(s).
- ACCW = continuous waves generated by an alternating-current plate supply (ACW).
- a.f. = audio frequency; in lists and tabular matter = A.-F.
- amp. = ampere.
- av. = average.
- ma. = milliampere.
- AVC = automatic volume control.
- B, b = susceptance.
- B = lines per square inch.
- \mathfrak{B} = flux density in gausses per square centimeter.
- C, c = capacity (farads).
- c.g.s. = centimeter-gram-second (absolute) system.
- cm. = centimeter.
- cond. = condenser.
- $\cos \phi$ = power factor $\frac{R}{Z}$; in text line R/Z .
- CW = continuous waves (unmodulated).
- d = distance; diameter; difference.
- db = decibel.
- d.c. = direct current.
- D/F = direction finder (radio).
- E = effective electromotive force; electrical pressure; volts.
- e.m.f. = electromotive force.
- e = instantaneous e.m.f.
- E, ϵ = electric-field intensity.
- \vec{E} = vector e.m.f.
- E_A = maximum or resultant e.m.f.
- \vec{E}_A = maximum or resultant vector e.m.f.
- E_c = grid-biasing voltage.
- E_{cn}, e_{cn} = counter e.m.f.
- E_o = grid-signal voltage swing.
- E.M.F., e.m.f. = electromotive force.
- E_0 = r.m.s. vector. $\left(\frac{E_A}{\sqrt{2}} \right)$
- E_p = vacuum-tube plate voltage.
- F = force (magnetomotive force in Gilberts); filament; modulation factor.

- f = frequency, cycles per second; filament; farad.
 f_c = cut-off frequency.
 fl = full load.
 G, g = conductance; vacuum-tube grid.
gal. = gallon(s).
 G_m = mutual conductance (micromhos).
 H = ampere turns.
 h = height.
h. = henry.
 \mathcal{H} = hysteresis; gilberts per centimeter.
 H = magnetic-field intensity; harmonic (pad type).
 H_2 = second harmonic (H_3 = third harmonic, etc.).
 I = effective current flow, strength of flow.
 i = instantaneous current.
 I_p = plate current.
ICW = interrupted continuous waves.
IF = intermediate frequency.
 $j = \sqrt{-1}$ (operating vector).
 K = numerical constant; dielectric constant.
 k = coupling coefficient.
kc./s. = kilocycles.
KVA., kva. = kilovolt-amperes ($E \times I \times 1,000$).
KW, kw. = kilowatt ($E \times I \times \cos \phi \times 1,000$).
 L = inductance.
 l = length.
log = logarithm.
 M = modulation percentage.
 M = mutual inductance; mutual coupling.
m. = meter (39.37 in.); modulation factor.
MA = milliammeter.
ma. = milliampere.
mc. or mc./s. = megacycle.
MCW = modulated continuous waves.
mf., mfd. = microfarad.
mmf. = micro-microfarad.
m.m.f. = magnetomotive force.
mh. = millihenry.
 μ h. = microhenry.
 $Mu = (\mu)$ = amplification factor or constant.
M. O. = master oscillator.
m.p. = maximum peak.
M.O.P.A. = master-oscillator power amplifier.
 N = total number of turns; turns ratio; number of field poles.
mv./m. = millivolts per meter.
m.p.h. = miles per hour.
mw. = milliwatts.
 n = number of turns.
 nl = no load.
NEMO = pick-up equipment remote from main-studio control room.
Osc. = oscillator.
 P = average electrical power (watts), $E \times I$; vacuum-tube plate.
P.A. = power amplifier.
 P, p = primary.

- p = instantaneous power.
P.D. = potential difference (volts).
 p.f. = power factor.
 Q = efficiency of a coil.
 Q, q = quantity of electricity in coulombs or ampere-hours.
QAVC = noise suppression control.
 R, r = resistance (ohms).
 \mathcal{R} = reluctance.
 R_o = resistance of output load or circuit.
 R_p = vacuum-tube alternating-current plate resistance.
RCA = Radio Corporation of America, Inc.
R.F., r.f. = radio frequency.
R.M.C.A. = Radiomarine Corporation of America, Inc.
RFA = radio-frequency ammeter.
 r.m.s. = root mean square.
 r.p.m. = revolutions per minute.
 r.p.s. = revolutions per second
 S, s = secondary; switch.
Sg, sg = screen grid.
 Sw. = switch.
 T, t = time in seconds.
 T-L = transmission line.
 V = vacuum tube.
 v = velocity.
VA, va. = volts \times amperes.
VM, vm. = voltmeter.
 VI = volume indicator.
 W = average electrical energy in joules or watt-hours ($E \times I \times t$).
 W = watts.
W.E. = Western Electric Co.
 w = instantaneous energy.
 X = reactance (reactive ohms).
 X_C = capacitive reactance (ohms).
 X_L = inductive reactance (ohms).
 X_M = mutual reactance (ohms).
 Y, y = admittance.
 Z = impedance (combined R and X in ohms).
 $|Z|$ = impedance magnitude (absolute value of)
 Z_i = impedance of input load or circuit.
 Z_o = impedance of output load or circuit.

SUBSCRIPT LETTERS

- A = applied.
 a = antenna.
 a.c. = alternating current.
 av = average.
 eff = effective value.
 fl = full load.
 m = maximum value.
 max = maximum value.
 min = minimum value.
 $ma.$ = milliamperes.

- $\mu\text{v./m.}$ = microvolt per meter.
 mv./m. = millivolt, per meter.
 nl = no load.
 i = input.
 o = output.
 P, p = primary.
 $p.r.$ = parallel-resonant.
 S, s = secondary.
 $s.r.$ = series-resonant.
 $r.m.s.$ = root mean square.

SPECIAL SIGNS

- θ = phase angle.
 λ = wave length.
 μ = permeability; amplification factor.
 μf = microfarad.
 $\mu\mu\text{f}$ = micro-microfarad.
 μh = microhenry.
 $\mu\text{v.}$ = microvolts.
 $\mu\text{v./m.}$ = microvolts per meter.
 $\mu\mu\text{w.}$ = micro-microwatt.
 ν = specific reluctance.
 π = 3.1416.
 ϕ = phase angle; power factor; flux (magnetic lines of force).
 ω = $2\pi f$; ohms.*
 Ω = ohms (standard); megohms.*
 \propto = is proportional to; varies as.
 \sim = cycle.
 ∞ = infinity.
 φ = angle.
 $| |$ = r.m.s. magnitude ($|E|, |Z_i$).
 \therefore = therefore.
 Δ = (delta) change, variation.
 \neq = is not equal to.
 \pm = plus or minus.

* It is the practice among many engineers to use ω to denote ohms and Ω to denote megohms. In this text ω denotes ohms and Ω denotes megohms. Examples: $10^\omega = 10$ ohms; $10^\Omega = 10$ megohms.

GREEK ALPHABET

Letters		Names	Letters		Names	Letters		Names
A	α	Alpha	I	ι	Iota	P	ρ	Rho
B	β	Beta	K	κ	Kappa	Σ	σ, ς	Sigma
Γ	γ	Gamma	Λ	λ	Lambda	T	τ	Tau
Δ	δ	Delta	M	μ	Mu	Υ	υ	Upsilon
E	ϵ	Epsilon	N	ν	Nu	Φ	φ or ϕ	Phi
Z	ζ	Zeta	Ξ	ξ	Xi	X	χ	Chi
H	η	Eta	O	\omicron	Omicron	Ψ	ψ	Psi
Θ	θ	Theta	Π	π	Pi	Ω	ω	Omega

COMMON PREFIXES

Abbreviation	Prefix	Numerical equivalent		Base and exponent	Characteristic of logarithms
meg. k.	mega	1,000,000		10^6	6. —
	kilo	1,000		10^3	3. —
		Fraction	Decimal		
deci.	deci	1/10	0.1	10^{-1}	— 1. —
c.	centi	1/100	0.01	10^{-2}	— 2. —
m.	milli	1/1000	0.001	10^{-3}	— 3. —
μ	micro	1/1,000,000	0.000001	10^{-6}	— 6. —
$\mu\mu$ (mm.)	micro-micro	1/1,000,000,000,000	0.000000000001	10^{-12}	— 12. —

PRACTICAL RADIO COMMUNICATION

CHAPTER I

DIRECT-CURRENT ELECTRICITY AND MAGNETISM

The *precise* nature of electricity is as undefined today as it was when electrical effects were first noticed centuries ago. Its manifestations, however, are now quite well understood, and from these, modern science has been able to construct a theory of electricity which satisfactorily explains most electrical actions.

It is upon this theory of electricity, known as the *electron theory*, that the art of radio is based. It is consistent, therefore, that a study of radio communication should begin with an exposition of the fundamental concepts of electricity as outlined in this chapter.

I. ELEMENTARY ELECTRICITY

1. Electric Charges and Forces.—It has been found that when certain bodies are subjected to friction, they possess a property of attraction or repulsion, depending upon the character of the materials used. This property has been defined as electricity or electrical energy, and is generally referred to as electrical charges showing differences in behavior. This is to say that, if one body acts in a definite manner, it is said to be positively electrified; and, if it acts in an opposite manner, it is said to be negatively electrified. It is merely an arbitrary expression of behavior to state that a body is positively or negatively electrified, charged, or excited. If a body does not react at all, it is said to be neutral, that is, the positive and negative charges are equal and therefore neutralize one another.

If a piece of sealing wax is rubbed with fur it is found that it acquires abnormal powers to attract other lighter bodies. This force of attraction is actually large enough to attract small bits of paper or deflect a suspended pith ball against the force of gravity.

If a pith ball is suspended from a silk thread, as illustrated in Fig. 1(a), and touched with a charged ebonite rod, the ball will acquire a negative charge. This is due to the ball receiving some negative particles of electricity from the rod, which consequently leaves it in an excessively

negative state. The ebonite rod will then repel the ball. If a charged glass rod is touched to another pith ball, some negative particles will pass from the ball to the glass, leaving the ball positively charged. The two pith balls thus oppositely charged and placed in the vicinity of one another will attract each other. Hence the law: *Like charges repel and unlike charges attract one another.*

Figure 1(a) may be used to show the phenomena of electrical forces existing between bodies and will serve as an illustration of a testing instrument for detecting small positive or negative forces. By applying negative particles to the suspended pith ball as previously stated, the instrument may be used as an effective electroscope to detect the presence and kind of charge in other bodies.

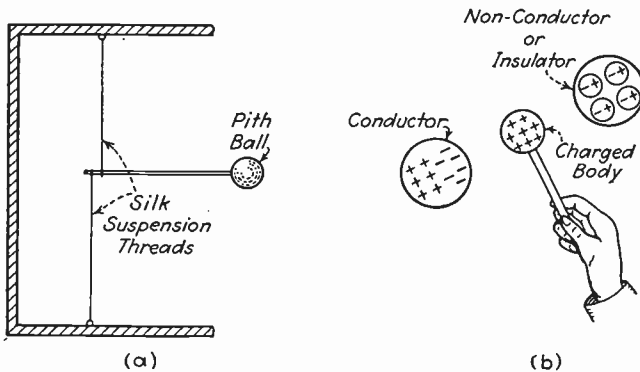


FIG. 1.—Phenomena of electrical forces between two bodies.

Any two dissimilar materials brought in contact become more or less charged with equal amounts of opposite kinds of electricity.

Uncharged or neutral bodies are believed to be composed of equal amounts of positive and negative particles of electricity. Thus, if a charged body is placed near it, the interior positive and negative particles of electricity are displaced in the manner illustrated in Fig. 1(b). It will be noted that the external charge attracts the unlike charge in the neutral material, and repels the like charge. If the material is a conductor, the negative charges are shifted to one end of the sphere and the positive charges to the opposite end. This is probably due to the fact that many of the negative particles move about freely in conductors, while in insulating materials or poor conductors the negative charges are bound, and are therefore not free to move about. In good insulating materials there are assumed to be few or no free charges, and yet attraction with charged bodies may result owing to the formation of the charge into small doublets. An approaching charge will attract these doublets as illustrated in Fig. 1(b).

It is this production of a state of strain between bodies that is called *electric or static lines of force*. The presence of these lines is fairly conclusive proof that all matter is a fundamental manifestation of electricity.

Matter in the general sense may be defined as anything that has weight, such as solids, gases, and liquids. Matter is believed to be composed of molecules which in turn are made up of one or more of the 92 known chemical elements. These elements of which the molecule is composed are known as *atoms*.

2. Protons and Electrons.—Matter, therefore, may be imagined as being made up, in the last analysis, of small individual bodies called *atoms*, which in turn are made up of small quantities of positive and negative charges called *protons* and *electrons*, respectively. The physical dimensions of these minute particles are now definitely known, particularly under conditions of rest. The proton has been calculated to have a mass 1.66×10^{-24} gram when at rest, and the electron to have a mass of 8.99×10^{-28} gram when at rest. These particles have a definite amount of charge about them which is called the *electrostatic field*. This field is a stationary one and is measured in units called *electrostatic units* or more recently *statcoulombs*. The statcoulomb, therefore, is an expression of the number of units of electricity or electric lines of force present when a body is at rest. The proton and electron both possess an indivisible charge about them of 4.77×10^{-10} electrostatic lines or statcoulombs.

From these dimensions it is quite apparent that the mass of an electron is an insignificant part of the total mass which is almost entirely the proton. This condition has now been definitely proved by the hydrogen atom, which incidentally is the smallest of all the atoms, and which contains one proton and one electron. In other words, the mass of a proton is substantially equal to the mass of the hydrogen atom which is calculated to weigh 1,840 times as much as an electron.

An atom is therefore a sort of solar system with a sun and planets and empty regions. These empty regions are of vastly greater dimensions than the imaginary sun and planets, so much so that in reality the greater part of a molecule, and in turn all matter, is unoccupied, although it may seem to us to be solid, gaseous, or liquid. In this imaginary solar system which makes up the atom we might conceive of the planets as the electrons and the sun as nuclear protons. It thus seems logical and according to scientific research findings that the weight of a body is dependent upon its atomic structure, and, in turn, that the activity in or outside of the atom is dependent upon the number of electrons around the nuclei.

When an atom has many electrons, it seems that they are arranged in successive orbits around the nucleus and revolve around it in either ellipses or circles. It is now definitely known that the chemical and

electrical properties of the atom depend almost entirely upon the outer ring of electrons. Copper and silver, for example, are classified as good conductors of electricity, probably due to the fact that the orbital arrangements of the electrons in these materials are such that the electrons in the outer orbit are more free to roam about than the outer orbital electrons of other materials. The electrons located in the inner structure of the atom are called *nuclear* electrons, and those in the orbits are called *orbital* electrons.

Figure 2 illustrates the arrangement of protons and electrons in a helium atom, showing both the nuclear and orbital electrons. It is now believed that nuclear electrons occur only as parts of a newly discovered entity called the *neutron*, the combination being located between the

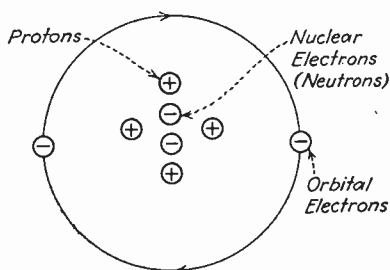


FIG. 2.—Arrangement of protons and electrons in a helium atom.

protons as compensating forces, since the repellent tendency of the four protons would otherwise cause them to fly apart.

Thus in all atoms, beginning with the hydrogen atom, which has one proton and one electron, up to the uranium atom, which has 92 protons and 92 electrons, the electrons are arranged in various orbits, and each element manifests certain chemical

and electrical phenomena dependent almost entirely upon the orbital motion of electrons, particularly in the outer ring.

It is now definitely known that in certain materials electrons are constantly changing from one orbit or orbital position to another. Under these conditions the material is constantly radiating energy into space which gradually causes its disintegration. This process takes place in those elements containing between 83 and 92 electrons in their atomic structure such as bismuth and uranium, respectively. Those elements of lower atomic numbers do not disintegrate by radiation unless they are subjected to a considerable rise in temperature under which condition electronic collisions increase and disintegration takes place.

Although physicists have also found other small entities to be included within the structure of the atom, the reader need not feel that a knowledge of their behavior is essential for an understanding of the flow of electrical current as treated in this text. In view of the frequent reference to these units in scientific literature, however, a brief summary of their characteristics will be given here.

The outstanding addition as a new particle of matter is the *positive electron*, which has been christened the *positron*. The existence of this particle was discovered by Dr. Carl D. Anderson in 1932. It is believed to consist of positively charged particles or corpuscles or rays. Its mass is equal to that of the electron but is of opposite polarity. Its charge, therefore, is like that of the proton although the latter has been computed to be approximately 2,000 times more massive.

The positron, according to Dirac, is born of radiation and possesses an extremely short life since it is immediately absorbed by surrounding matter and dies only to give birth to new radiation. The precise manner of its annihilation and rebirth has been deduced from the fact that when it strikes a free electron, both particles are annihilated and give rise to what are known as *photons* or *gamma rays* which travel in opposite directions with a total energy of a million *electron volts*.

It must not be believed, however, that the positron is a preexisting particle of matter in an atom such as the proton, neutron, and electron; it is rather a particle entirely born of radiation.

3. Potential.—The term *potential* is commonly used to express the level of any point in space. In an electric field, for example, the potential of a point has the same meaning as has potential energy to a point in the gravitational field of the earth. That is, the potential of any point above the surface of the earth (ground) may be defined as the amount of energy that is required to raise a unit mass of matter from the earth to that point. More energy or work is required to raise it to a greater height from the earth, and therefore the potential energy required is said to be greater. When a mass is raised to a point at the limit of the gravitational field of the earth or, in other words, to an infinite distance, then its potential with relation to the earth is at its maximum.

It is generally regarded that the outermost boundary of an electric field has an absolute zero potential and that, therefore, this *absolute potential* of any point in a field may be defined as the quantity of work required to bring unit mass, if the field is gravitational, from infinity to that point. Similarly the *absolute electrical potential of a point is defined as the work necessary to bring a unit-positive charge from an infinite distance to the point*. If the field is due to a positive or negative charge, the potential of any point in the field is either positive or negative, since work must be performed by an external force, and by a resisting field. Therefore, the potential energy of the unit positive or negative charge will be increased when it is moved closer to the charge which is repelling it.

In an electrical circuit we say that a current flows when an electrical potential difference exists between the two ends of the circuit. Hence, if two oppositely charged bodies, are connected by a conductor, a current will flow from the negative to the positive point or from the point of low potential to the point of high potential. For example, if the + and the - charges of Fig. 3 are connected by a conducting wire, the electrons will pass from the negatively charged body to the positively charged body, and in doing so will carry one end of the lines of force with them, finally shortening the lines until they disappear. Hence the strain between the bodies is said to be relieved and the potential difference drops to zero.

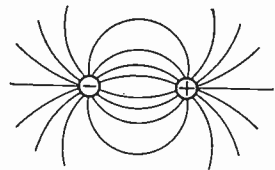


FIG. 3.—Electric field between two oppositely charged bodies.

The potential difference (*PD*) between any two points may be defined as the amount of work required to move a unit-positive or unit-negative charge from one point to another.

From this it may readily be seen that whenever work is being done by moving charges in an electric field a difference of potential must exist between the two charged bodies. In other words, differences in potential must exist wherever there is an electric field, and an electric field must exist whenever there are differences of potential.

If it is desired to find the electrical potential at a point near a charged body we may do so by the following reasoning.

In Fig. 4 the point *A* is a charged unit with a certain number of statcoulombs which will be called *Q* units of electricity. To obtain the

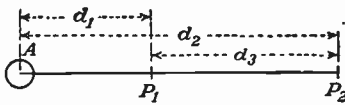


FIG. 4.—Illustrating the difference of potential between a charged body *A* and various points *P* at various distances *d*.

difference of potential, *PD*, between the point *p*₁ at a certain distance *d*₁ from the charged body *A*, and the point *p*₂ at a distance *d*₂ from the charged body *A*, the strength of the electric field at the point *p*₁ may be obtained by taking the ratio of the electric units to the distance squared, or $\phi_1 = Q/d_1^2$; similarly, at the point *p*₂, by the ratio $\phi_2 = Q/d_2^2$. Now, since the potential difference is the work required to move unit-positive charge from *p*₂ to *p*₁, that is, over distance *d*₃, and if the strength of the field is equally distributed between *p*₂ and *p*₁, the amount of work done is found by multiplying the strength of the field by the distance *d*₃. However, since this force or strength of the field varies inversely as the square of the distance from *A*, the average force can only be obtained by taking the geometric average of the two forces *p*₁ and *p*₂, thus, $\sqrt{\frac{Q}{d_1 \times d_2}}$. The work or the potential difference may then be obtained by taking the product of the average strength of the field times the distance. Hence,

$$PD = \frac{Q}{d_1 \times d_2} \times (d_2 - d_1) = Q \left(\frac{1}{d_1} - \frac{1}{d_2} \right) \quad (1)$$

If it is required to obtain the absolute potential *E*₁ of the point *p*₁, the point *p*₂ must be moved to an infinite distance from *A*. Hence, the potential difference *E*₁ at the point *p*₁ will be equal to the ratio of the statcoulombs to the distance *d*₁, or *Q*/*d*₁. From this it may readily be seen that the potential due to a charged point varies directly as the charge and inversely as the distance from the charge. A generalization may now be stated as *Coulomb's law*, as follows. *Force is inversely proportional to the square of the distance between electrically charged particles.*

4. Free Electrons in Solid Conductors.—In order to understand electric phenomena, as associated with the flow of current in a cir-

cuit, some idea as to the atomic structure in this relationship must be understood.

Various phenomena point to the existence of roving electrons in a solid conductor called *free* or *conduction electrons*. It is believed that these electrons are constantly moving from atom to atom in a conductor with a movement which resembles an elliptical motion. That is to say, some electrons leave their respective atoms and move toward other atoms inside the conductor without being forced to do so by the application of an exterior electrical pressure.

These electrons apparently do not travel from end to end in a conductor, as they do in an electrical circuit when some external potential is applied. They probably move from atom to atom and, like little planets, swing around their nuclei.

Thus we must imagine them to have left their orbits in the atoms of the solid and to be moving about in the solid in a sort of haphazard way. Consequently the atoms that have lost an electron are positively charged and, even though they are vibrating because of material temperature, are not free to move about the solid. This is due to the effect of molecular forces in the vicinity of these atoms which hold them in their definite positions. It is apparent here that the free electrons moving about may again recombine with atoms, but in the same material their number per cubic centimeter is assumed to be almost universally constant; that is, when one electron recombines, another electron becomes free. At temperatures above 450°C. these free electrons actually force themselves through the surface of the metal and radiate energy into space. At normal metal temperatures the free electrons that take part in the transfer of electricity are sometimes called *conduction electrons*.

5. **Electric Current.**—The modern conception of an electric current flow in a conductor, as accepted by physicists, is based upon the theory of conduction electrons; in other words, that an electric current may be defined as a progressively moving electron, or electrons. This may give the impression that an electric current is actually flowing in a wire having both ends open, since free or conduction electrons are actually in motion. That is true, but, since the electrons do not actually flow from one end of the wire to the other end unless they are forced to do so by the application of some external potential, they are not considered “flowing” from the electric-current point of view. However, as soon as the two ends of a conductor are connected to a battery or generator this movement of electrons begins and continues around the entire circuit in a definite direction, depending upon the polarity of the battery or generator, as long as the difference of potential is applied.

Motional Velocity of Electrons.—When a potential is applied to two ends of a wire conductor to form an electric circuit, the electrons move around the entire circuit from the negative terminals of the battery or

generator to the positive terminals. This is contrary to the earlier belief that a current always flows from the positive terminal through the conducting medium to the negative terminal. This latter conception is still maintained in many textbooks only to conform to the conventional standard of direction.

When it is said that a current is flowing from the positive to the negative terminals of a battery or generator this merely expresses the *conventional* flow of an electric current. But it is, nevertheless, true that electrons are moving in the opposite direction, that is, from the negative terminal of the potential supply through the metallic conductor toward the positive terminal. Hence, the current flow in a metallic conductor must always be looked on as an electron drift toward the positive terminal from the negative terminal of the battery or generator.

It is important to note that the electrons do not move at a high rate of speed through the metallic conductor when the potential is applied. That is to say, the progressive motion of the electrons in a metallic conductor is very slow. The electrical impulse or signal set up by the electron movement, however, is very fast and attains the velocity of an electric wave through the conductor or surrounding medium. This is analogous to a heavy freight train which when starting moves as a whole very slowly, but on which the pulling force or energy is transferred very quickly with a jerk from the engine to the last car.

The slowness of the actual electron movement is due to the fact that the free electrons are constantly colliding with the atoms in the metal, consequently retarding their progressive motion. On the other hand, it must not be thought that the actual velocity of the electrons is low. If it were not for the constant collisions with the atoms of copper in the wire, the electrons would actually reach velocities many millions of times as great as their normal drift. However, since the copper atom or molecule is many thousand times as great as that of the electron, the numerous collisions prevent a rapid progressive motion. The electrons, therefore, bound and rebound at tremendous velocities in the copper wire, always moving toward the positive electrode; thus producing an electron drift through the entire conductor.

It can readily be seen that the electron would acquire tremendous velocities if it did not collide with the relatively larger masses of atoms or molecules in the wire. Proof of this tremendous accelerating action is seen in hot metals. The heating of a body tremendously increases the atomic and molecular agitation, and the electrons may be accelerated to such a high degree that they are actually forced out through the surface of the metal. Under these conditions electrons are known to have reached tremendous velocities in the order of thousands of miles per second.

Summarizing, it is believed that the existence of free electrons in the conductors is the real reason for the manifestation of an electric-current flow in an electric circuit when a pressure is applied. It is therefore also to be assumed that any material or body which is so atomically constructed as to be devoid of or deficient in free electrons is a poor conductor. This seems to indicate that materials classified as

insulators are apparently lacking in free-electron movement within their atomic and molecular structure, and that conductors offer resistance to current flow in inverse proportion to the number of free electrons present in their molecular structures.

The reader must bear in mind that the only true flow of current in solid conductors is the electron drift through the circuit from the point of negative potential to the point of positive potential, the potential in electric circuits being provided by batteries or generators.

In gases or liquids the current flow is in both directions. This action will be described in detail later.

6. Electromotive Force.—In order to maintain a steady flow of current in an electric circuit, there must be a constant pressure and a suitable path through which the current may flow. The pressure in electrical circuits is called the *electromotive force* or e.m.f. In hydraulic systems the pressure is referred to as a certain number of units or pounds per square inch. Electrically this pressure is expressed in units called *volts*, which are the units used to denote the difference of potential or electromotive force. Hence, when the e.m.f. of a battery or generator is mentioned, the reference is to its pressure as being a certain number of volts.

The electron viewpoint on the theory of potential difference is of far greater importance than the more generally accepted hydraulic analogy, because it expresses more definitely the actual conditions which take place in an electric circuit. For example, let us refer to Fig. 5 which represents a simple electric circuit of copper wire connected to the terminals of a battery or generator E .

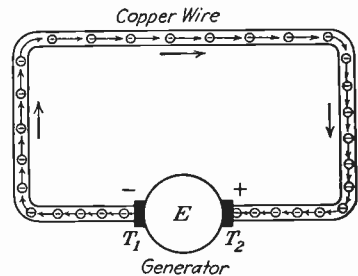


FIG. 5.—A simple electrical circuit showing current flow.

The circuit must be visualized as a continuous conductor completed through the medium of the generator winding between the terminals T_1 and T_2 . When the generator is set in motion, a generation of e.m.f. takes place (to be explained in Chap. XIV), and an e.m.f. is exerted between the terminals T_1 and T_2 . This establishes a difference of potential which will make one terminal of the generator positive (+) and the opposite terminal negative (-). Now, since the entire circuit forms a continuous path through the conductor and the generator winding, the force exerted by the generator in motion causes the free electrons in the copper-wire conductor to be moved in the direction of the arrows, as indicated in the diagram. The electrons progressing around the conductor in the direction of the positive terminal are said to be moving from a point of low potential to one of a higher potential. The reason for this progressive motion is probably due to the fact that electrons in the generator winding have been displaced during the time in which the generator is in motion. This condition may be hypothesized by the illustration in Fig. 5, where E again represents the terminals of the generator.

When the generator is put in motion, a displacement of electrons in the generator winding will force a number of free electrons into the conducting wire at the negative terminal T_1 . The displaced electrons now moving in the direction of the arrow into the conductor will repel other free electrons in the wire beyond them. Hence there will be a succession of forces transmitted through the entire wire by virtue of these displaced electrons. Thus the continuous displacement of electrons in the generator armature during its motion causes all the free electrons in the conductor to move around the circuit from T_1 to T_2 producing what is generally known as an *electric current*.

It is important to note that the same number of electrons must pass any given point in the conductor in a given time so long as the current remains constant. This can be readily visualized if one imagines a long pipe filled with billiard balls which are receiving a steady series of pushes at one end causing the balls in the entire pipe circuit to move. Hence, the movement of the balls will be the same at every section of the circular tunnel (pipe) regardless of the section at which the pushes are produced. Electrically this means that a current flow is the same at any portion of a simple circuit.

7. Current Strength and Quantity.—The rate of flow of water through a pipe is measured as so many gallons a second, which expression includes a definite quantity of water and a unit of time. The distinction between the terms *rate of flow* (strength) and quantity must be carefully understood in order to comprehend the practical application of these terms to electric circuits. For example, at the rate of 1 gal. of water per second, there might be 3,600 gal. of water delivered to a tank in 1 hr., thus distinguishing the rate of flow from the quantity, and naming it. Electrical quantity is measured in *coulombs*. In referring to the relation between quantity and rate of flow in an electric circuit, the electrical term for rate of flow or strength must be used. This unit is called the *ampere*. When one practical unit of quantity of electricity (coulomb) flows continuously every second, the rate of flow, or the strength, is said to be 1 amp., or, if 2 coulombs flow continuously every second, the strength of the current would be 2 amp., and so on. Hence, it can be readily seen that the current in amperes is independent of the length of time the current flows in a given circuit, regardless of whether it flows for a fraction of a second or for hours.

To find the total quantity of current flowing through a circuit in a given time, multiply the amperes by the time (in seconds),

$$Q = I \times t \quad (2)$$

Example: If an incandescent lamp requires a current of $\frac{1}{4}$ amp. to maintain a steady brilliancy, what quantity of electricity would be consumed if the lamp is lighted 2 hr.?

$$2 \text{ hr.} = 60 \times 60 \times 2 = 7,200 \text{ sec.}$$

Formula and Computation:

$$Q = I \times t = \frac{1}{4} \times 7,200$$

$$Q = 1,800 \text{ coulombs}$$

To find the average current strength (in amperes) when the quantity and the time are known:

$$I = \frac{Q}{t} \quad (3)$$

Proof:

$$Q = 1,800 \text{ coulombs}$$

$$t = 7,200 \text{ sec.}$$

$$I = \frac{1,800}{7,200}$$

$$= 0.25 \text{ or } \frac{1}{4} \text{ amp.}$$

To find the time (in seconds) required for a given quantity of electricity (in coulombs) to pass a certain point of a circuit:

$$t = \frac{Q}{I} \quad (4)$$

Proof:

$$Q = 1,800 \text{ coulombs}$$

$$I = \frac{1}{4} \text{ amp.}$$

$$t = \frac{1,800}{\frac{1}{4}}$$

$$= 7,200 \text{ sec.}$$

8. Electrical Resistance and Conductivity of Materials.—When water flows through a pipe, the resistance it meets with depends directly on the length of the pipe, its diameter, and general conditions, such as bends and roughness. Similarly, when a difference of potential is applied to an electric circuit, a current flows through the circuit, and the amount of opposition (resistance) offered to the flow is proportional to the resistivity of the wire, which in turn depends upon the character of the material through which the current flows, its length, diameter, and temperature.

It will be remembered that the materials having less free or conduction electrons in them are classified as poor conductors or resistances. Hence alloys which have less free electrons in them than copper or silver are higher resistance materials, and the progressive flow of electrons will be relatively less; consequently we say that the current through any such material will be reduced. All metals, therefore, have a certain amount of resistance or conductivity which depends upon their atomic structure.

When pure metals are heated, through excessive current flow or otherwise, the electron acceleration in the material is greatly increased. This results in a greater number of electron impacts with the atoms, consequently reducing the progressive motion of electrons in the conductor. Heated wires, therefore, have greater resistance than cool wires, depending upon the amount of heat and the character of the metal.

It has been found that the electron flow in certain alloys when placed in an atmosphere approaching *absolute zero* temperature is practically unimpeded. That is to say, owing to the fact that all molecular motion in heatless bodies is reduced to zero, practically *no resistance* is offered to the progressively moving electrons when an electromotive force is applied. The *absolute zero* temperature, commonly known as the Kelvin temperature, abbreviated K., is 460 degrees below zero on the Fahrenheit scale or 273 degrees below zero on the Centigrade scale. Hence it may readily be seen that at ordinary temperatures there is an ever present molecular agitation which consequently introduces an opposition or resistance to the flow of free electrons.

The resistance of carbon varies inversely with temperature. In other words, the resistance of carbon decreases as the carbon gets hot. Therefore, the *hot* resistance of a carbon lamp is less than its *cold* resistance. On the other hand, the resistance of a tungsten or other metallic-filament lamp, including metallized carbon, increases with temperature. The hot resistance of a tungsten filament is about 12 times its cold resistance.

The resistance of certain (most) alloys, such as German silver, does not vary appreciably with temperature.

The resistance of electrolytes decreases as the temperature rises.

The following table illustrates the relative resistance and conductance values of different pure metals and alloys when operating under the temperatures indicated.

TABLE I.—RELATIVE CONDUCTANCES AND RESISTANCES OF CONDUCTORS

Material	Resistance of a 1 cir.-mil.-ft. (ohms)		Relative conductivity (as compared with copper)	Relative resistance (as compared with copper)
	30°F.	75°F.		
Silver, pure annealed.....	8.83	9.67	108.6	0.925
Copper, annealed.....	9.59	10.5	100.0	1.00
Copper, hard drawn.....	9.8	10.75	97.8	1.02
Aluminum, 95 per cent pure	16.03	17.7	59.8	1.67
Zinc, pure.....	34.6	37.9	27.7	3.60
Iron wire.....	58.7	65.1	16.2	6.17
Nickel.....	74.1	85.13	12.95	7.72
Steel wire.....	81.16	90.15	11.6	8.62
Brass.....	43.31	45.4	22.15	4.51
Phosphor-bronze.....	51.0	51.8	18.8	5.31
German silver.....	127.0	128.7	7.5	13.32

An examination of the above table shows that the resistance of alloys is very much greater than that of the pure metals. This is a characteristic property of alloys, which is taken advantage of in the preparation of wires of high specific resistance. Even a slight trace of another metal, which by itself may be a good conductor, has an enormous effect on the resistance; hence, copper used for electrical purposes has to be exceptionally pure.

The resistance of any material is expressed in *ohms*. (The electrical sign for the ohm is usually shown thus ω .) This is the term used to express the resistance of an electrical circuit in ohms per unit length. For example, the resistance of 1,000 ft. of copper wire which has a diameter of $\frac{1}{10}$ in. (No. 10 B. and S. gage) is about 1 ohm, although a piece of iron wire of the same length and cross section has a resistance of

about 6 ohms, and a similar piece of German silver wire has a resistance of about 17 ohms.

9. Ohm's Law.—In any circuit through which there is a flow of current we must obviously have all of the three following factors present: (1) the *pressure* or *potential difference* (volts) which causes the current to flow; (2) the opposition or "resistance" (ohms) which must be overcome to produce a current flow; (3) the *current strength* (amperes) which can be maintained in a circuit as a result of the pressure overcoming the resistance and thus causing a flow. In any circuit there is always a definite relation between these three units, and, therefore, the value of any one unknown factor may be calculated when the values of the other two are known. The law governing these calculations is known as Ohm's law.

RULE 1: *The current strength in any circuit is equal to the e.m.f. applied to the circuit, divided by the resistance of the circuit.*

$$\text{Current} = \frac{\text{pressure}}{\text{resistance}}$$

or

$$I = \frac{E}{R} \quad (5)$$

Example 1: If an electric lamp having a resistance of 55 ohms is connected across a potential of 110 volts what current will flow through the lamp?

$$I = \frac{E}{R} = \frac{110}{55}$$

$$I = 2 \text{ amp.}$$

The current strength in any circuit increases or decreases directly with the increase or decrease in the potential, when the resistance in the circuit is assumed to be constant. If the pressure is constant, the current will increase as the resistance is decreased, and decrease as the resistance is increased.

In other words the current might be said to vary directly with the e.m.f., and inversely as to the resistance.

Example 2: If in the above problem the voltage is increased to 220 volts, how many amperes will flow through the lamp?

$$I = \frac{E}{R} = \frac{220}{55}$$

$$I = 4 \text{ amp.}$$

RULE 2: *The amount of e.m.f. required to maintain a certain current strength in a circuit in which the resistance is known is equal to the product of the current strength and the resistance*

$$\text{Pressure} = \text{current} \times \text{resistance}$$

or

$$E = I \times R \text{ (written } IR\text{)}. \quad (6)$$

Example 3: How much pressure must be applied to a circuit to cause 5 amp. to flow if the resistance is 30 ohms?

$$E = I \times R = 5 \times 30$$

$$E = 150 \text{ volts}$$

The pressure varies directly with the current and resistance values. For example, if it is desired to send a greater current through the same resistance, a greater amount of pressure must be applied to the circuit, or, if the same current is to be passed through a greater resistance, then a greater pressure must be applied.

RULE 3: *To find the value of resistance required to be inserted in any circuit, so that a given current will flow under a known pressure: the resistance is equal to the pressure to be applied, divided by the current strength to be maintained:*

$$\text{Resistance} = \frac{\text{pressure}}{\text{current}}$$

or

$$R = \frac{E}{I} \quad (7)$$

Example 4: A certain resistance passes a current of 7 amp. through a circuit at a pressure of 35 volts. What is the value of the resistance?

$$R = \frac{E}{I} = \frac{35}{7}$$

$$R = 5 \text{ ohms}$$

When a constant pressure is desired, its resistance must be cut in half if the current is to be doubled, or, if there is to be a constant current maintained in which the pressure is doubled, then the resistance must also be doubled.

A simple method of remembering the Ohm's-law application to direct-current measurements is illustrated in Fig. 6.

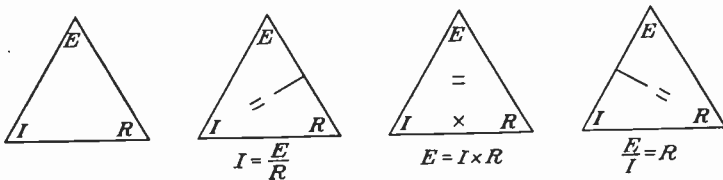


FIG. 6.—A simple method of remembering Ohm's law.

For example, if the voltage and amperage in an electrical circuit are known, the pyramid diagram may be applied as follows: Place the finger over the unknown quantity, that is, the resistance R . This leaves the remaining letters E and I for the voltage and current, respectively. Thus the letter E being above the letter I simply means that the current I is divided into the voltage E to give R . If the voltage, for example, was found to have been 100 volts and the current 2 amp., then by the application of the pyramid explanation, the resistance must be 50 ohms.

Similarly, if it is desirable to find the current I if the voltage E and the resistance R are known, then, by the same application, the current can be determined by placing the finger over the unknown quantity I . For example, if the voltage E is 100 volts

and the resistance R is 50 ohms, then, by application of the pyramid, it will be found that the current I will be 2 amp. Similarly, if the resistance R in an electric circuit is 50 ohms and the current I 2 amp., then, by application of the pyramid, the unknown quantity E can be determined by placing the finger over the E . Thus multiplying IR we shall find the voltage E to be 100.

10. Simple Electric Circuits.—Ohm's law shows that, for a given voltage, the lower the resistance the larger will be the current, and the higher the resistance the smaller the current.

Circuits are divided into three classes:

1. Series circuits.
2. Parallel circuits.
3. Series-parallel or parallel-series circuits.

a. Series Circuits.—Figure 7 shows a simple series circuit in which may flow a steady current. This is called a series circuit because the current flows in one continuous path. Furthermore, this current is the same at any point in the circuit. An ammeter placed at any point would give the same reading. The ammeter is a device for measuring current strength in amperes.

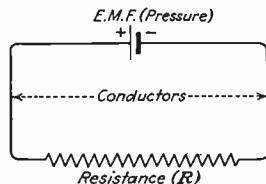


FIG. 7.—A simple series circuit.

Now, as each of these parts will have resistance, each will have what is known as an IR drop, which is always directly proportional to the resistance. Hence, each IR drop represents a certain amount of e.m.f. expenditure in each part of the circuit, and the differences of potential produced by the source must equal the total e.m.f. expended. In other words, the input must equal the output plus losses. This can be readily seen if the three parts used in all series circuits are considered.

1. The inside or internal part of the e.m.f. source, that is, batteries, generators, etc.

2. The leads which connect the apparatus to the source, that is, the lines.

3. The apparatus itself.

Thus, if each one of these parts has a certain amount of resistance, and the same current is flowing through them, each will have an IR drop which is proportional to its resistance. Therefore the IR drop in the source is called the *internal drop*, and that in the line is called the *line drop*. Hence, since each IR drop represents an expenditure of e.m.f., the total expenditure of e.m.f. in the circuit will be the sum of the e.m.f.s. expended in each part, and, obviously, the difference of potential produced by the source must equal the total e.m.f. expended. Therefore, the supply e.m.f. must always supply a higher voltage than is required at the source terminals. The source voltage is referred to as

a definite *no-load* voltage when no current is being delivered to the external circuit, and as a *load* voltage whenever current is being forced through the circuit. Obviously, there will be a certain voltage drop whenever current is drawn from a source, the amount, of course, being proportional to the resistance of the circuit.

It is seen, therefore, that any source of e.m.f. having a high internal resistance will not permit a large current to flow in the external circuit, because even a small current will cause an internal expenditure of voltage equal to the total voltage of the source and, therefore, no voltage will be available at the terminals for external use.

Figure 8(a) shows a series circuit with more than one resistance unit connected in series. Remembering that the same amount of current passes through every part of the circuit, the total e.m.f. will be the sum of all the IR drops. Thus, the equivalent resistance (a

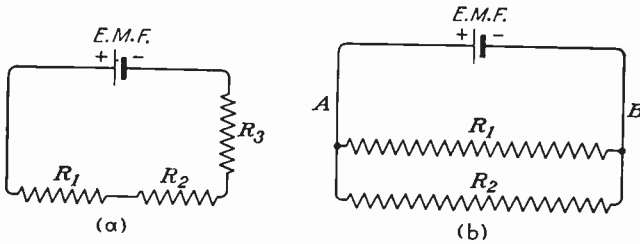


FIG. 8.—Series and parallel circuits.

resistance having a total value of the three) in a series circuit is equal to the sum of the resistances of the individual units (assuming the wires to have a negligible resistance), thus:

$$\text{Total effective resistance, } R = R_1 + R_2 + R_3, \text{ etc.} \quad (8)$$

Referring to Fig. 8(a), assume each resistance to have 5 ohms resistance; then what would be the effective circuit resistance?

$$R = R_1(5) + R_2(5) + R_3(5) = 15 \text{ ohms}$$

The total resistance in the circuit, therefore, would be equivalent to a circuit containing one resistance of 15 ohms, and so on. Thus, the current would be equal to the total e.m.f. divided by the total resistance, or

$$I = \frac{E}{R}$$

An important point to remember in series circuits is that, for a given voltage, the current at every point in the circuit is inversely proportional to the total resistance of the circuit.

b. Parallel Circuits.—A parallel circuit is one in which there are two or more parts connected between two points in a circuit. Figure

8(b) shows a simple parallel circuit consisting of two resistances R_1 and R_2 connected between two points a and b of any circuit. These resistances are assumed to have a value of 10 and 20 ohms, respectively, and to be connected across a potential of 100 volts.

How many amperes are flowing through each resistance and what is the total amperage being drawn from the supply line?

$$R_1 = 10 \text{ ohms}$$

$$R_2 = 20 \text{ ohms}$$

Then, by Ohm's law as applied to a series circuit, the current flowing through R_1 is found by dividing 100 into 10 or $I = E/R$. Thus

$$I = \frac{E}{R_1}$$

$$I = 10 \text{ amp.}$$

Then, the current flowing through R_2 will be found in a like manner to be

$$I = \frac{E}{R_2}$$

$$I = 5 \text{ amp.}$$

and so on, regardless of the number connected in parallel. The total current flowing in the exterior circuit, or in other words the combined current, is then found by adding the current flow of each part, thus

$$I = I_1 + I_2$$

or

$$I = 10 + 5 \text{ amp.}$$

$$I = 15 \text{ amp.}$$

In parallel circuits, therefore, the total current flowing through any parallel combination can always be found by applying Ohm's law to each branch and then adding the current value of each branch, as in the last example.

It can be seen, from the above statement, that the total current I is greater than can possibly be obtained through any one branch providing the impressed e.m.f. is kept constant.

From this, it is quite obvious that the joint resistance of any parallel combination is less than the resistance of any one of the branches.

For example, if three resistances of 10, 20, and 30 ohms are connected in parallel, the total effective resistance will be less than the smallest resistance. The effective resistance, therefore, can be accurately found by the formula

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}} \quad (9)$$

where $R_1 = 10$ ohms.

$R_2 = 20$ ohms.

$R_3 = 30$ ohms.

Hence

$$R = \frac{1}{\frac{1}{60} + \frac{1}{60} + \frac{1}{60}} = \frac{1}{\frac{3}{60}} = \frac{60}{3} = 20$$

$$R = 5 + \text{ ohms}$$

When this formula is applied to *two* resistances in parallel, it becomes

$$R_{eff} = \frac{R_1 \times R_2}{R_1 + R_2} \tag{10}$$

c. Series-parallel Circuits.—A series-parallel or parallel-series circuit comprises a combination of series and parallel parts. Figure 9 illustrates

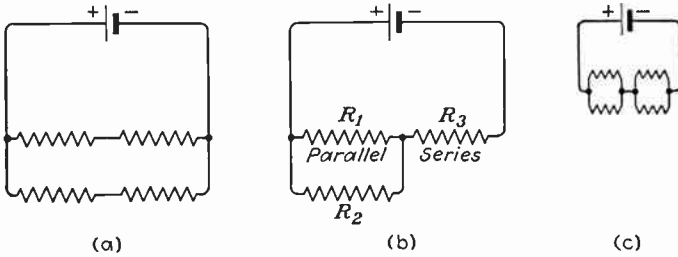


FIG. 9.—Series-parallel and parallel-series circuits.

three types of series-parallel and parallel-series circuits. Once again Ohm's law must be applied to every part of the circuit. In analyzing more complicated circuits of this type, however, it will be necessary first to reduce each parallel combination to its equivalent series resistance before combining it with the remainder of the circuit. Then, the circuit-resistance voltage drop and current flow can be determined by Ohm's law.

For example, if R_1 and $R_2 = 10$, and $R_3 = 5$ ohms, then by reducing the parallel combination R_1 and R_2 by the parallel formula, we find the resistance to be 5 ohms. Then, by adding R_3 to the result of R_1 and R_2 , we find the total effective resistance to be 10 ohms. The same application can be made to the two other forms of series-parallel and parallel-series circuits.

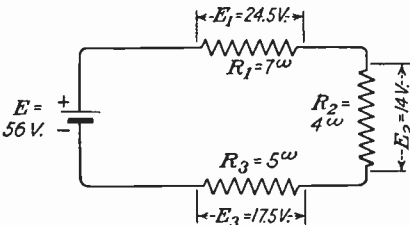


FIG. 10.—A simple series circuit containing three resistances in series.

d. Circuit Applications.—There are, however, many variations of simple series, parallel, and parallel-series circuit problems, such as are illustrated in some of the following diagrams.

Figure 10 is a simple series circuit containing three resistances in series. The total resistance of the entire circuit is readily determined by adding the three resist-

ances; hence the effective resistance, $R_{eff} = R_1 + R_2 + R_3 = 16$ ohms. By the use of Ohm's law $I = E/R$, we find that the current flow in the circuit is 3.5 amp. Since the current flow in a series circuit is the same at any portion of the circuit, 3.5 amp. must flow through each resistance. Now, if it is desired to determine the voltage or potential difference across any one of the resistances, we simply apply Ohm's law, $E = IR$. Thus, if it is desired to determine the voltage across E_2 , multiply IR_2 , or $3.5 \times 4 = 14$ volts. It can readily be seen that the resistances R_1 and R_3 are providing a potential drop of 24.5 and 17.5 volts, respectively. In other words, these two voltage drops deducted from 56 volts leave a balance voltage of 14 across R_2 , which proves that the sum of the voltage drops across each resistance must equal the applied e.m.f., or in this case 56 volts.

Figure 11 illustrates a series-parallel combination. In this type of circuit the parallel combination of 10 and 5 ohms must first be reduced to its effective resistance by the formula on parallel resistances,

$$R_{eff} = \frac{R_1 R_2}{R_1 + R_2}$$

or an effective resistance of $3\frac{1}{3}$ ohms. Since the effective resistance of R_2 and R_3 is connected in series with R_1 , the total circuit resistance may then be obtained by adding

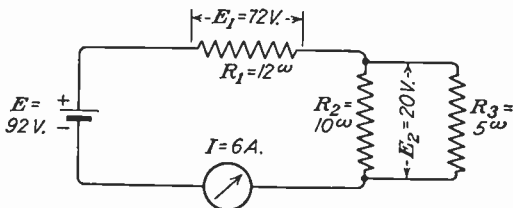


FIG. 11.—A series-parallel combination.

the two values $R_1 + R_{eff}$, or $3\frac{1}{3} + 12$, which is a total resistance of $15\frac{1}{3}$ ohms. The current flow in the circuit may again be obtained by Ohm's law $I = E/R = 92/15\frac{1}{3} = 6$ amp.

To prove this example we find that the voltage across $E_1 = IR_1 = 6 \times 12 = 72$ volts and $E_2 = IR_{eff} = 6 \times 3\frac{1}{3} = 20$ volts. Hence, $E_1 + E_2 = 72 + 20 = 92$ volts, which is the value of the applied e.m.f.

It will be noted that the voltage pressure across R_2 and R_3 is the same. This condition holds true in all parallel circuits since it is assumed that the connecting wires have a negligible resistance. In parallel combinations, therefore, it is a simple matter to determine the current flow through each branch by the ratio $I = E/R$.

In the preceding example we find the current flow by this ratio through $R_3 = I = E/R = 2\frac{2}{5} = 4$ amp.; through branch $R_2 = E/I = 2\frac{2}{5} = 2$ amp., or a total of 6 amp. flowing out of the parallel combination.

With this form of reasoning it is possible to obtain the unknown values of any series or series-parallel combination.

For example, Fig. 12 illustrates a series-parallel combination in which the applied e.m.f. and total current flow are given, but one of the parallel resistances is unknown. To find the value of this resistance proceed as follows.

Determine the amount of current passing through R_2 , since the difference between I through R_2 and the total flow of 4 amp. must be the amount passing through the unknown resistance. When the current flow through R_3 is found, the resistance may be determined by Ohm's law $R = E/I$.

The voltage across the parallel combination E_2 must be 16 volts, since the resistance R_1 is carrying a current of 4 amp.; hence $E_1 = IR_1$, or 8 volts. Thus, if the resistance R_1 has a voltage drop across it of 8 volts, the remaining voltage across the parallel combination must be 16 volts. Hence the current flowing through $R_2 = I_2 = E_2/R_2 = 16/12 = 1\frac{1}{3}$ amp. and, since the total current flow is 4 amp., the current flow through R_3 must be $I - I_2 = 4 - 1\frac{1}{3} = 2\frac{2}{3}$ amp. Then $R_3 = E/I = 16/2\frac{2}{3} = 6$ ohms.

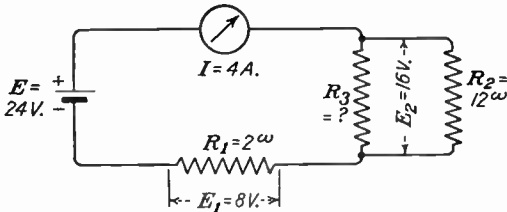


FIG. 12.—Another series-parallel combination.

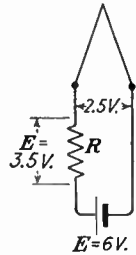


FIG. 13.—Resistance in series with filament.

A practical application of the simple series circuit is illustrated in Fig. 13 in which a vacuum tube is designed for an operating e.m.f. of 2.5 volts at which voltage the tube will draw 1.75 amperes. What value of resistance must be inserted to operate this tube at the required voltage if the applied e.m.f. is 6 volts?

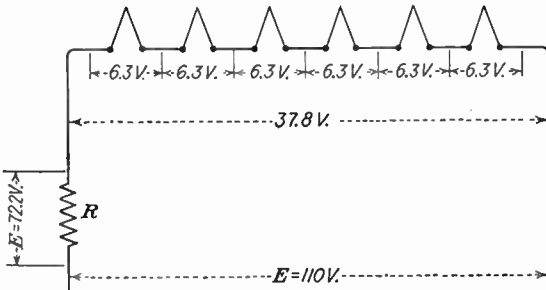


FIG. 14.—Filaments in series.

Since the voltage across the tube terminals must be 2.5 volts, the resistance will be required to drop the voltage from 6 to 2.5 volts, a drop of 3.5 volts. Hence, if a current flow of 1.75 amp. passes through the resistance and filament, under the assumption that R is correct, we may readily determine the value of resistance to obtain the desired result by the ratio $R = E/I = 3.5/1.75 = 2$ ohms.

Similarly, if a number of low-voltage filaments are to be operated from a high-voltage supply system, the voltage must be decreased to the desired operating voltage of the tubes. For example, six tubes are

to be operated from a 110-volt line and each to require a potential of 6.3 volts. At this potential each tube will draw 0.3 amp. What value of resistance must be inserted to satisfy this condition? Figure 14 illustrates the circuit arrangement.

If each tube requires a potential of 6.3 volts, then the effective potential of the six tubes connected in series must be 6×6.3 volts, or 37.8 volts. Hence, it will be necessary to drop the voltage from 110 to 37.8 volts, or a total drop of 72.2 volts.

The total current flow through R and all the filament resistances will be 0.3 amp. (current flow in a series circuit is the same at any portion

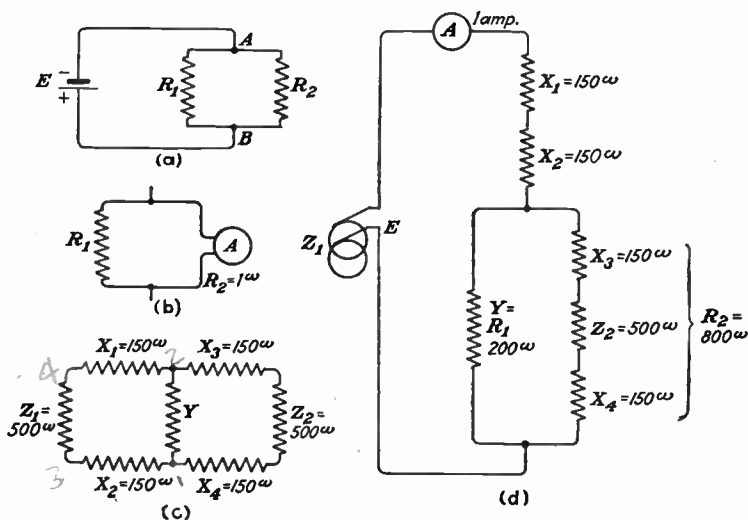


FIG. 15.—Circuits illustrating the shunt law.

of the circuit), if the value of R is assumed to be correct. Hence the correct value of R to effect this condition is again obtained by the ratio $R = 72.2/0.3 = 240\frac{2}{3}$ ohms.

11. The Shunt Law.—When an e.m.f. is applied to a network consisting of two resistances in parallel, the current divides itself between the two paths in inverse proportion to the resistance they offer to its flow. Any branch of a circuit which by-passes another branch is called a *shunt* with regard to the branch it by-passes.

Thus, in Fig. 15(a):

$$\frac{I_1}{I_2} = \frac{R_2}{R_1}$$

and, therefore,

$$I_2 = I \left(\frac{R_1}{R_1 + R_2} \right) \tag{11}$$

where I_1 and I_2 are the currents in R_1 and R_2 , respectively, and I is the total current flowing from A to B . The factor $\frac{R_1 + R_2}{R_1}$ is called the *multiplying factor* of the shunt.

Example 1: An ammeter A reads 0 to 5 amp. It is desired to design a shunt S to be used with the meter to cause it to read 0 to 25 amp. What should the resistance of the shunt be, if the resistance of the meter is 1 ohm? The circuit is shown in Fig. 15(b).

Solution:

$$\frac{R_1}{R_2} = \frac{I_2}{I_1}$$

$$\frac{I_2}{I_1} = \frac{5}{20} = \frac{R_1}{R_2} = \frac{1}{4}$$

then

$$R_1 = \frac{1}{4}R_2 = \frac{1}{4} \times 1 = \frac{1}{4} \text{ ohm}$$

The multiplying factor of the shunt R_1 is, therefore, found by the equation

$$\frac{R_1 + R_2}{R_1} = \frac{0.25 + 1}{0.25} = \frac{1.25}{0.25} = 5$$

So the shunt multiplies the current-carrying capacity of the parallel circuit R_1, R_2 by 5. Whereas the meter would register full deflection with 5 amp. flowing through the circuit without the shunt, it registers full deflection with 25 amp. when by-passed by the shunt, a multiplication of five times.

Example 2: It is desired to know what current will flow through a load resistance Z_2 of 500 ohms connected to the output of an H pad (attenuation network) as shown in Fig. 15(c). The equivalent of this circuit is shown as Fig. 15(d). The ammeter A reads 1 amp.

Solution:

$$\frac{R_1}{R_2} = \frac{I_2}{I_1} = \frac{200}{800} = \frac{I_2}{I_1}$$

Therefore

$$I_2 = I \left(\frac{R_1}{R_1 + R_2} \right) = I_2 = 1 \left(\frac{200}{1000} \right)$$

$$= 1 \times 0.2 = 0.2 \text{ amp.}$$

If 0.2 amp. flows through R_2 , then 0.2 amp. must flow through Z_2 , as the current in all parts of a series circuit is constant.

12. Kirchhoff's Laws.—In the simple electric circuits explained, the equivalent resistance and the currents in the various branches are readily calculated by Ohm's law, but in cases where complex combinations involving cross connections are employed the solution cannot be derived by that law.

Equations for the solution of such problems can easily be written down by using Kirchhoff's two laws. These are not "laws" in the same sense as Ohm's law; they are merely useful rules which show how Ohm's law is to be applied.

Rule 1: In any branching network of wires the algebraic sum of the currents in all the wires meeting at a point is zero.

The term *algebraic sum* means that, if the currents are flowing from the point, they are taken with a negative sign; and if flowing to the point they are taken with a positive sign.

The rule may also be expressed: *The algebraic sum of the potential drops around every closed circuit is always equal to zero.*

This law then simply states that no electricity accumulates at any point in a circuit.

A simple illustration of Kirchoff's first law is seen in Fig. 16.

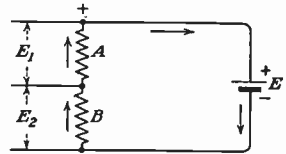


FIG. 16.—A circuit illustrating Kirchoff's law.

The law states that if the voltage is measured across the resistance A, and then B, the sum of these two voltage drops must be the same as the applied voltage E. In other words, there are two reactions in the circuit, namely, the one which causes the current to flow through the resistance, due to the applied difference of potential, and the other, the resistance reaction in the opposite direction, which tends to stop the current. This is an important point in calculating problems in which several e.m.fs. are applied in one circuit, some of which may be in opposite directions.

Figure 17 illustrates one of these conditions. In this circuit the total impressed e.m.f. acting in the circuit is $E + E_2 - E_1$, illustrating that there are two forces

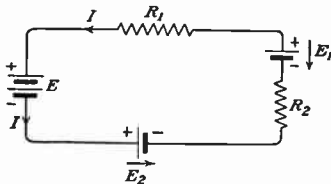


FIG. 17.—A circuit illustrating Kirchoff's first law.

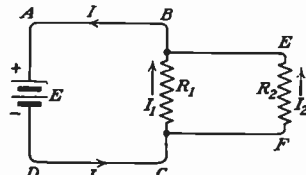


FIG. 18.—Another circuit for Kirchoff's first law.

supplementing each other, and one opposing the two. Hence, if the impressed force is made equal to the sum of the IR drops in the circuit, we may write

$$E + E_2 - E_1 = IR_1 + IR_2 \tag{12}$$

or

$$E = IR_1 + IR_2 + E_1 - E_2 \tag{13}$$

It is therefore simply an expression, that the sum of the IR drops in the circuit must be equal to the algebraic sum of the various e.m.fs. acting in the circuit.

In circuits where the current flow is apparent, such as in Fig. 18, the solutions for the problem are easily obtained by Kirchoff's first law.

For path ABCD:

$$E = I_1R_1 \text{ or } E - I_1R_1 = 0$$

For path *A E F D*:

$$E = I_2 R_2 \text{ or } E - I_2 R_2 = 0$$

It can be readily seen that circuits containing e.m.fs. in different directions can be very easily solved if we call all the voltages which would cause the current to flow in one direction positive, and those which cause it to flow in the opposite direction negative. Thus all the positive and negative voltages are added up algebraically. Similarly all the *IR* drops due to current in one direction are called positive, and those *IR* drops due to current in the opposite direction are called negative.

Rule 2: In any closed path in the network, the algebraic sum of the IR products is equal to the e.m.f. acting in that path.

The rule may also be expressed: *The algebraic sum of the currents at any junction of conductors is always zero.*

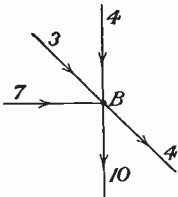


FIG. 19.—Illustration of Kirchhoff's second law.

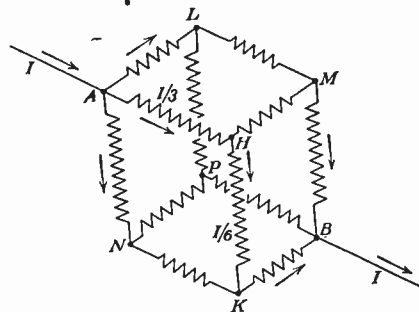


FIG. 20.—A complex electrical circuit.

This condition is illustrated in Fig. 19 in which case all the electrons moving into the junction *B* are also flowing away from the junction, or, in other words, there can be no accumulation of electrons at any one point. Hence, the algebraic sum of all the currents at *B* must be zero.

Complex electrical circuits, such as illustrated in Fig. 20, may be readily solved by application of Kirchhoff's second law. Assume a network of resistances made up in the form of a cube in which each edge has an equal resistance of a value *r*. To find the effective resistance of the entire mesh between the points *A* and *B* proceed as follows.

Assuming *I* to be the total current, and *E* the potential difference between the points *A* and *B*, then, by equality, the current divides itself into three equal parts at the point *A* and again reunites at the point *B*. Hence, currents in *AH* and *KB* are each *I/3*. Also at *H* there are two equal paths for the current *I/3* and, therefore, the current in *HK* must be *I/6*.

The voltage between the points *A* and *B* is the sum of the voltages across *AH*, *HK*, and *KB*. Therefore,

$$E = \left(\frac{I}{3} \times r\right) + \left(\frac{I}{6} \times r\right) + \left(\frac{I}{3} \times r\right) = \frac{5}{6}Ir$$

but if R_{eff} = the equivalent resistance, then $E = IR$

Therefore,

$$IR = \frac{5}{6}Ir \text{ or } R_{eff} = \frac{5}{6} \times r$$

Another form of complex network is illustrated in Fig. 21. Here a number of resistances are connected in series and parallel combinations with a shunt resistance connected in between them.

This system is relatively more difficult to analyze owing to the additional shunt resistance between the arms. To find the effective resistance of the entire network proceed as follows:

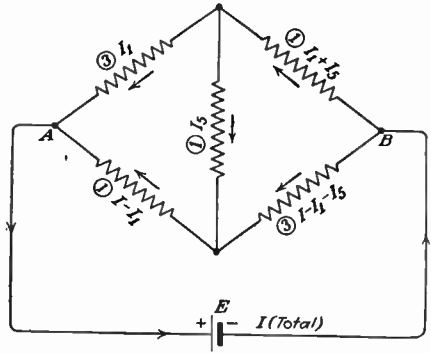


FIG. 21.—A simplified version of a bridge network.

Total $I = I_1 + (I - I_1)$ etc., as shown in the following solution.

Voltage equations

Current equations

(1) $3I_1 - I_s - (I - I_1) = 0$

$4I_1 - I_s = I$

(2) $I_s + I_1 + I_s - 3(I - I_1 - I_s) = 0$

$5I_s + 4I_1 = 3I$

(3) $I - I_1 + 3(I - I_1 - I_s) = E$

$4I - 4I_1 - 3I_s = E$

(4) $3I_1 + I_1 + I_s = E$

$4I_1 + I_s = E$

Substituting (1) in (3):

(5) $4I - 4I_1 - 12I_1 + 3I = E = 7I - 16I_1 = E$

Eliminating I_s between (1) and (2)

$$-5I_s + 20I_1 = +5I$$

$$5I_s + 4I_1 = 3I$$

$$\hline 24I_1 = 8I$$

(6)

$$3I_1 = I$$

$$I_1 = \frac{I}{3}$$

Substituting (6) into (5):

$$7I - 16\left(\frac{I}{3}\right) = E$$

$$E = 7I - 5\frac{1}{3}I = 1\frac{2}{3}I = \frac{5}{3}I$$

hence since $R = \frac{E}{I}$

$$R_{eff} = 1\frac{2}{3} \text{ ohms}$$

Problems of this character should be carefully checked by assuming a certain voltage E applied to the network at the points A and B .

Proof: If $E = 20$ volts, then I (total) at a resistance of $1\frac{2}{3}$ ohms would be 12 amp. since

$$I = \frac{E}{R} = \frac{20}{1\frac{2}{3}} = 12 \text{ amp.}$$

Hence, by computing the IR drops across each 3-ohm resistance, and each 1-ohm resistance on the outside, we find the sum of the IR drops to be equal to the applied e.m.f. (see Fig. 22).

13. Conductance.—It has been seen that the opposition offered to the flow of electrons in a circuit is called the resistance, consequently all materials possessing high-resistive properties are poor conductors; that is to say, the conductivity or conductance of such materials will be poor. Conversely, if the opposition to the flow of electrons in a material is low, the conductance is good.

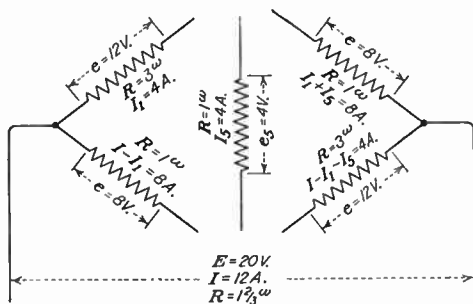


FIG. 22.—Diagram for proving complex circuit computation.

The conductance of a circuit is, therefore, the reciprocal of resistance and is expressed by the ratio

$$G = \frac{1}{R} \quad (14)$$

where G represents the symbol for conductance.

Since G is the reciprocal of the resistance unit, namely, the ohm, the term conductance of a circuit is expressed as so many mhos (ohms spelled backward). For example, if a wire has a resistance of 1 ohm, the conductance will be 1 mho; if of 2 ohms resistance, $\frac{1}{2}$ mho; 8 ohms resistance, $\frac{1}{8}$ mho; $\frac{2}{3}$ ohm resistance, $\frac{3}{2}$ or $1\frac{1}{2}$ mhos, etc. If in Ohm's law, conductance is used instead of resistance the law would read

$$I = EG \quad (15)$$

$$E = \frac{I}{G} \quad (16)$$

$$G = \frac{I}{E} \quad (17)$$

14. Energy, Force, Work, and Power.—*Energy* may be defined as the ability or capacity to do work. *Force* is one of the factors of work and has to be exerted through a distance to do work. *Work* is reckoned as the product of the force and the distance through which it has been applied. Work is done when energy is expended or when force over-

comes resistance. *Power* is the time rate of doing work or the rate at which energy is dissipated or transferred in the circuit.

15. Electrical Power.—Assuming a current of I amp. flowing for a certain number of seconds t in a circuit at a potential of E volts, the amount of electricity (coulombs) flowing in the circuit would then be a result of the product of $I \times t$ in which I is expressed in ampere-seconds. From this the energy expended in the circuit is measured by the product $E \times I \times t$ and is called the *energy units* or *ergs* and is expressed in watt-hours.

The practical unit of e.m.f., the volt, is essentially made up of 10^{-8} units of e.m.f. called *absolute units of e.m.f.* These small units are not used in practical work but serve to indicate how certain fundamental principles are definitely concluded. Similarly, the ampere is also expressed in *absolute units of current* so that we say 1 amp., $\frac{1}{10}$ of an absolute unit of current, and hence the energy expended in heat is obtained by the equation

$$\text{Energy expended} = E \times 10^{-8} \times I \times \frac{1}{10} \times t \text{ ergs} \tag{18}$$

Therefore,

$$W = E \times I \times t \times 10^7 \text{ ergs, or } E \times I \times 10^7 \text{ ergs per second} \tag{19}$$

Hence the rate at which energy is expended in heat in an electrical circuit is obtained by the product $E \times I$. This is called the power expenditure and is referred to as so many power units, called *watts*. Thus, if an e.m.f. of 1 volt causes a current of 1 amp. to flow through a resistance of 1 ohm, the power expended in the circuit will be 1 watt; hence, 1 watt = 10^7 ergs per second. This is generally referred to as so many watts active in 1 sec. or *joules*. Thus 1 watt acting for 1 sec. is called a *joule*.

Power Expressions.—Power = EI watts. Now, since

$$E = IR, \quad \text{and} \quad I = \frac{E}{R}, \tag{20}$$

it follows that

$$\text{Power} = I \times R \times I = I^2R \text{ watts} \tag{21}$$

or

$$\text{Power} = E \times \frac{E}{R} = \frac{E^2}{R} \text{ watts} \tag{22}$$

The three expressions are numerically identical, and consequently one should use the form which is most convenient.

In saying that the three expressions are numerically identical, it must be understood that E is the voltage which forces the current through the resistance R . For example, let us assume that a current I flows between two points in a circuit. If the potential difference, PD , between

these points is represented by E volts, then the rate of power expenditure in that portion of the circuit will be given by the product EI as so many watts. If, however, the expression I^2R is used to denote the rate of power expenditure, it must be understood that this equation refers only to the actual waste of power by conversion into heat. For example, if an ordinary resistance is connected across a supply voltage, all of the power is expended in heat. But, if a piece of apparatus such as a motor is connected across the voltage, some of the power is wasted in heat, while some of the power is converted into useful work. Under these conditions the product I^2R gives merely the power wasted in heat, and not the useful or working power. Hence it may be seen that as more power is wasted in I^2R less useful power will be available for the operation of the motor. Although the expression I^2R may generally be looked upon as wasted power, this is not entirely true since the power used up as heat in an electric heater is really useful power.

When mechanical work is transformed into heat or heat into work, the amount of work done is always equivalent to the quantity of heat.

16. Load Matching.—Perhaps one of the most important fundamentals in electrical engineering for the maximum development of power in a load is the proper design of the load device to match the resistance or impedance of the generating device itself. Whether the generating source be a generator, a chemical cell, or a vacuum tube, it must possess internal resistance or impedance. Consequently, if a load

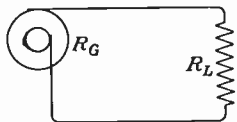


FIG. 23.—Circuit to illustrate load matching.

is connected to any of these generating devices, maximum power can be absorbed by the load only if the two devices are of the same resistance or impedance value.

For example, let us assume the generator (R_G) in Fig. 23 to have an internal resistance (due to its winding) of 1 ohm. We shall assume that, at open circuit or no-load condition, the voltage developed at the terminals of the generator is 3 volts. Now, if the load resistance (R_L) is 1 ohm, the total power dissipation in the generator winding R_G and the load resistance R_L will be 4.5 watts, since

$$I = \frac{E}{R_G + R_L} = \frac{3}{1 + 1} = 1.5 \text{ amp.}$$

Hence,

$$P_{\text{total}} = E \times I = 3 \times 1.5 = 4.5 \text{ watts}$$

Each resistance, therefore, dissipates an equal amount of power or 2.25 watts. Now, if the load resistance is doubled so that $R_L = 2$ ohms, the total current flow will be

$$I = \frac{E}{R_L + R_G} = \frac{3}{1 + 2} = 1 \text{ amp.}$$

Hence

$$P_{\text{total}} = E \times I = 3 \times 1 = 3 \text{ watts}$$

The power dissipated by the load resistance, therefore, will be

$$P_L = I^2 \times R = 1 \times 2 = 2 \text{ watts}$$

and that dissipated by the generator resistance

$$P_G = I^2 \times R = 1 \times 1 = 1 \text{ watt}$$

Now, if the load resistance R_L is small compared with R_G , for example 0.5 ohm, the current flow will be $I = \frac{E}{R_G + R_L} = \frac{3}{1 + 0.5} = 2$ amp. Hence the total power dissipated in the two resistances will be

$$P_{\text{total}} = E \times I = 3 \times 2 = 6 \text{ watts}$$

Thus the power dissipated in R_L will be

$$P_L = I^2 \times R = 4 \times 0.5 = 2 \text{ watts}$$

while that dissipated in the generator winding R_G will be

$$P_G = I^2 \times R = 4 \times 1 = 4 \text{ watts}$$

Although a total of 6 watts is expended in the circuit, only 2 watts are available at the load. Hence, 4 watts are wasted in heat in the generator winding. This relation may be more clearly illustrated if we assume the load resistance R_L to be of zero resistance value. Under these conditions, the power available at the load points (the short-circuited resistance) would be zero, and consequently all of the power would be wasted in heat in the generator winding itself.

17. E.M.F. Produced by Chemical Action.—If two dissimilar substances are placed apart in certain chemical solutions, a difference of potential will be found to exist between them. Of these substances, copper and zinc, or carbon and zinc, immersed in a solution of sulphuric acid and water, constitute the most commonly used primary cells, while lead dioxide and sponge lead in a diluted solution of sulphuric acid, or a nickel-iron combination in an alkali solution, constitute the two types of secondary cells.

In order to make an electric cell it is necessary to provide two electrodes one of which is at a higher potential than the other, in consequence of which a current will flow between these two electrodes if they are connected through an external circuit. In order to raise the potential of one electrode over the other, the two electrodes are immersed in an electrolyte which may take one of several forms. For storage batteries a diluted solution of sulphuric acid is used, while in dry cells it is a mixture

of sal-ammoniac and zinc chloride. As the action of the sulphuric acid and water is the simplest to study, this action will be explained.

Let us examine the action in a primary cell (Fig. 24) made up of one copper and one zinc electrode immersed in a dilute solution of sulphuric acid, H_2SO_4 , the acid being dissolved in 3 parts of water.

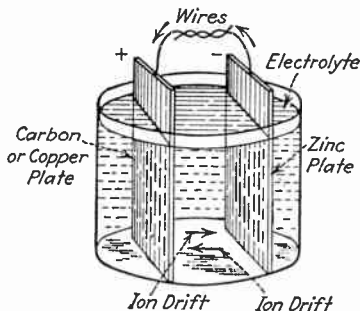


FIG. 24.—A simple primary cell.

a. When the acid is so dissolved the electrolyte breaks up into a simple positive radical or ion of hydrogen, H^+ , and a compound negative radical or ion of sulphur and oxygen (SO_4^-). There is then the electrolyte in which, figuratively speaking, there are two radicals or ions in free suspension. The hydrogen ion carries a positive charge, and the sulphuric ion carries a double negative charge.

b. Owing to the chemical action of the electrolyte on the zinc electrode, atomic disintegration takes place which causes atoms to leave the surface of the zinc and combine with the negative ions (SO_4^-) in the electrolyte, forming zinc sulphate which dissolves in the water. During the operation of the primary cell the zinc electrode is, therefore, literally eaten away, and it is obvious that such a cell cannot be recharged. It can be renewed by the insertion of a new zinc electrode and a renewal of the electrolyte.

c. The loss of positive charges from the surface of the zinc leaves a predominance of electrons, negative charges, and the zinc becomes negatively charged. If a circuit is provided from the zinc electrode to the copper electrode, a current will flow which will consist of a movement of the negative electrons from the zinc through the wire, over to the copper.

Hydrogen does not unite with copper (or carbon); thus the positive ions of the electrolyte are simply attracted toward the copper plate, which is negative with respect to hydrogen. This results in the formation of hydrogen gas on the plate. Here the gas clings in the form of bubbles until its lightness causes it to rise to the top of the electrolyte and escape into the air. However, under certain conditions the hydrogen bubbles do not rise rapidly enough to clear the surface of the copper electrode, and thus the *internal resistance* of the cell is increased. This formation of hydrogen gas on the surface of the positive electrode is called *polarization*, and unless counteracted, becomes a distinctly limiting factor in the life of a primary cell. To counteract polarization, oxygen must be introduced into the battery to combine with the hydrogen atom, forming water and thus preventing polarization.

d. Because of its porous nature and its ability to absorb oxygen and thus prevent polarization, carbon is preferred to copper as the positive terminal of the primary cell. The perfection of the so-called *air cell* was permitted by the discovery of a special form of carbon which literally breathes in oxygen, thus delaying polarization and making possible a primary cell with a life of a thousand hours or more.

e. The production of an e.m.f. in a secondary cell is essentially due to the same type of action just described for the primary cell. Planté developed the first commercially practical storage battery about the year 1860. To this day the lead-acid storage battery is essentially the same as that developed in those early days, except, of course, for mechanical improvements in making the plates and containers. The positive plate of this cell consists of a grid covered with lead dioxide, and the negative plate consists of pure sponge lead. The electrolyte is a diluted solution of sulphuric acid.

f. If an external load circuit is connected to this cell so a current can flow, the action is as follows. Atoms from the surface of the positive plate (lead dioxide, PbO_2) combine with the positive hydrogen ions, H^+ , liberated in the electrolyte giving the plate a positive charge and forming lead monoxide which later reacts with the sulphuric acid to form lead sulphate. The positive plate therefore becomes filled with lead sulphate as the discharge progresses.

At the same time, the negative sulphuric ions (SO_4)⁻ combine with the atoms of the negative plate (Pb) forming lead sulphate on the surface of the plate. Positive charges having been removed from the surface of the sponge-lead plate to form lead sulphate, an excess of negative charges remains on the plate, and it acquires a negative charge or potential. The formation of lead sulphate on the surfaces of both plates is called sulphation and tends to impede the combining of the atoms of the active material of the plates with the ions liberated by the electrolyte, and the *internal resistance* of the cell is increased. As the sulphation continues, the internal resistance increases and the voltage produced drops. Lead sulphate, unlike zinc sulphate, is non-soluble in water.

g. If a current is forced through the cell in a direction opposite to the discharge current, the chemical action of discharge may be reversed, that is, the lead sulphate on the surface of the plates can be decomposed and its constituent ions forced back to their original position in the electrolyte and active materials of the plates. When all of the sulphate has been decomposed and the elements restored to their original state, the cell is said to be charged. It is obvious, of course, that the charging current must be greater in value than the terminal voltage of the cell; otherwise, the cell will continue to discharge and this charging current must be maintained at this greater value until the charging process has been completed.

The chemical action in a nickel-iron-alkali cell is very different from that of the lead-acid type; it is much more complex but essentially the results are the same; an e.m.f. is produced owing to one plate being at a higher potential than the other. The details of this action will be described later, in the section on the Edison Cell (see Sec. 408).

18. Directions of Current Flow.—Current flow in a metallic conductor is a flow of free electrons toward the positive terminal of the conductor. In the early days of electricity it was assumed and agreed that the *conventional* flow of electric current should be a positive flow, that is, a flow from the positive toward the negative. As pointed out in Sec. 5 on Electric Current this conventional flow is still referred to as *the* flow of electric current whereas it is only the *conventional flow*, the actual electronic flow being in the opposite direction in the most commonly considered conductor, that is, in the metallic conductor. When the path is a gaseous one, as in certain types of vacuum tubes, the electrons liberated by electronic collision are attracted toward the positive plate or terminal, and any positive ions are attracted toward the filament which is the negative terminal. This constitutes a flow of current in two directions: one a positive flow, and the other a negative flow; one electronic and the other consisting of positive ions. This positive flow is usually so small that it can be neglected. Its presence, nevertheless, has been proved by laboratory experiments.

The flow of current in an electrolyte consists of a flow of negative ions toward one electrode and a flow of positive ions toward the other electrode. This constitutes a flow of current in two directions, and the flow of negative electricity is not proportionately greater than the flow of positive electricity.

The directions of current flow in an electrolyte are not of much practical importance. Furthermore, as the flow of current in a vacuum tube is almost wholly an electronic flow, the positive flow, when present, can usually be neglected. In metallic conductors the flow is entirely electronic which makes it apparent that for all practical purposes *current flow is electron flow* which is always toward the positive potential.

A clear understanding of what constitutes a current flow in metallic, gaseous, and electrolytic conductors should clear up any questions as to direction of current flow. It should be understood that any statement to the effect that current flows from positive to negative refers only to a flow in a gaseous or electrolytic conductor, or to the conventional flow of current in a conductor, in which the actual flow is electronic and, therefore, from negative to positive. Bear all this in mind when studying diagrams on which the direction of current flow in a metallic conductor (wire) may be indicated by arrows which, if pointing from positive to negative, indicate *conventional* flow, not *actual* electronic flow.

19. Local Action.—The presence of impurities in the plates of an electric cell sets up small local currents, owing to the action of the electrolyte on the two dissimilar metals. These impurities are embedded in the active material of the plates and may be carbon, iron, or arsenic, or other foreign matter. The local currents set up by these impurities within each plate reduce the effectiveness of the plate for its intended purpose and act as a constant drain on the cell, reducing the mass of the affected plate by actual disintegration in a primary cell or by sulphation in a lead-acid secondary cell. This waste of energy and active material is called *local action*, as it is set up locally in the cell regardless of external conditions. It is counteracted in part by using the purest obtainable material in the manufacture of the cell.

20. The Voltage of a Cell.—The e.m.f. produced by a cell depends only upon the kind of material used for the electrodes and the character of the electrolyte in which these electrodes are immersed. The size or shape of the electrodes do not affect the value of e.m.f. produced; but these factors do affect the capacity of the cell as will be seen later.

The voltage on open circuit is the potential difference between the positive and negative electrode. When the circuit is closed, the value of voltage is decreased by the amount of e.m.f. lost in overcoming the internal resistance of the cell and in forcing the current to flow through it. Therefore, as a dry cell gets old, and polarization increases its internal

resistance, its terminal voltage drops proportionately. A voltage drop takes place also in the lead-acid cell as the plates sulphate and the internal resistance of the cell is increased.

It is for this reason that an open-circuit voltage reading is of little value in determining the potential charge in a cell. This electrical condition is, of course, affected by the *internal resistance* of the cell and must be measured under a condition which will vary with a change in the internal resistance. Such a condition is one in which a current flows through the cell, and a voltage reading is valuable only when taken with the cell discharging in a normal way.

21. The capacity of a cell is usually measured in ampere-hours, which may be expressed in unit form as the number of hours a cell will deliver a given current to a final terminal voltage specified by the manufacturer and usually, on storage batteries, is indicated on the name plate.

The ampere-hour capacity of a lead-acid storage cell is proportional to the area, thickness, porosity, and arrangement of the active material, the quantity and specific gravity of the electrolyte, the temperature under which the battery is operating, and the rate of discharge. It is evident, therefore, that the ampere-hour capacity of a cell is very largely dependent upon its design characteristics.

As dry cells are not generally required to deliver much current they are not rated in ampere-hours, although, of course, every electric cell whether a primary or secondary cell may be rated in this way if desired. In specifying dry cells it is generally more important, after the number required is known, to specify that newly made cells be supplied, as the best of dry cells deteriorate even when standing on a shelf on open circuit, and their life rarely exceeds 15 months even on open circuit. It may be said then that the capacity of a dry cell depends, in addition to the factors enumerated above, on its age.

22. Connection Methods.—It is evident that, other factors being constant, the voltage of a cell is dependent only on the character of the material constituting its plates, and an increase in voltage over that delivered by one cell can be had only by adding together the e.m.fs. of more than one cell. This is called *connecting cells in series* and is graphically shown in Fig. 25(a).

In order to increase the capacity of a battery, that is, to increase the load it can carry, it is necessary to increase the area affected by the electrolyte. As an electric cell is usually available in only a limited number of cell sizes, it is convenient, from a practical standpoint, to increase the capacity of the battery (collection of cells), when dry cells are used, by connecting two or more cells in parallel as shown in Fig. 25(b). The parallel scheme is also referred to as connecting in multiple. For example, if it is said that three cells, or lamps, are connected in multiple, it is the same as saying they are connected in parallel.

It is possible by utilizing a combination of series and parallel connections to get any value of voltage or current-delivering capacity desired. Such a scheme of connecting is called the series-parallel method and is shown in Fig. 25(c).

23. The Resistances of a Cell.—While reference has been made before in this chapter to the internal resistance of electric cells, the subject is important enough to justify this summation.

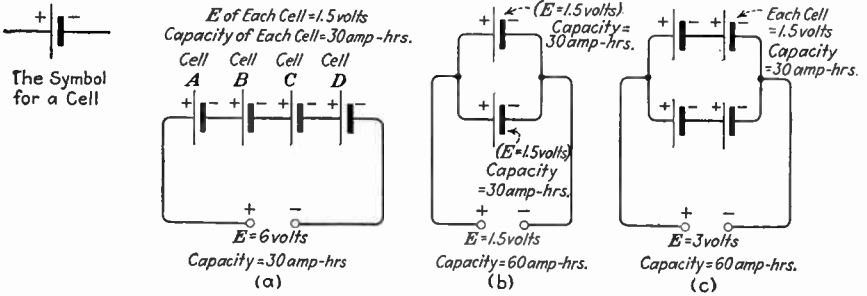


FIG. 25.—Various ways of connecting cells.

Mathematically, the *resistance* of a cell, R , is the ratio of the difference between the voltage on open circuit and the voltage under load divided by the current under load and is expressed by the equation

$$R = \frac{E_1 - E_2}{I}$$

where E_1 is the no-load voltage reading and E_2 the full-load voltage reading. I is the current reading made simultaneously with E_2 .

Physically the *internal resistance* is the electrical resistance of the path the current must follow within the cell in flowing to one terminal of the cell or the other. This includes the resistance of the electrolyte and that of any porous electrode material. It also includes the resistance in a primary cell caused by polarization, which is an extremely variable factor being caused by the accumulation of gases on the surface of the cathode electrode, and which is accompanied by a counter e.m.f. set up by the contact of the hydrogen gas on the cathode material. In a lead-acid secondary cell it includes the resistance caused by sulphation, which is also a variable factor.

The *internal resistance of a cell varies* directly with the area and spacing of the active materials exposed to the electrolyte, as the cross section and length of the electrolytic conductor are thus regulated. It is obviously affected by the strength or density of the electrolyte, as the character of the conducting material is thus affected. It is also evident that the contact between the electrolyte will vary with the porosity of the active materials which accounts in part for the higher internal resistance of a discharged cell compared with a charged cell and the consequent lower

voltage of the discharged cell. Finally, it varies with the temperature of the electrolyte, as the conductivity of all conductors is affected by temperature.

It is seen then that there is a fine distinction between the so-called resistance of a cell which means *total* resistance, and the internal-resistance value which usually refers to a new dry cell or a fully charged storage battery. The factor *internal resistance* affects the efficiency of a cell, whereas the factor *total resistance* is useful in gaging the state of charge of a cell.

24. Dry Cells.—The production of electricity by a dry cell is governed by the action taking place under similar circumstances in a wet primary cell, the only difference being in the method of applying the electrolyte.

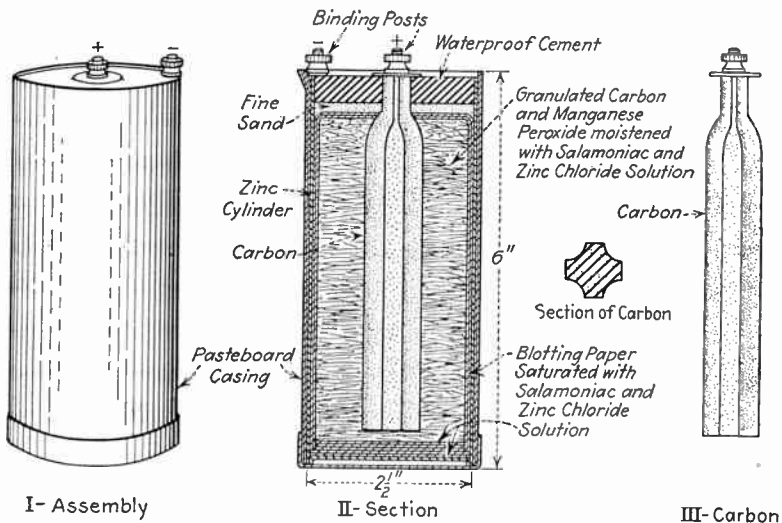


FIG. 26.—Construction of dry cell.

The most common form of individual dry cell is that known as the Standard 6, being 6 in. high and 2 1/2 in. in diameter. Its construction is shown in Fig. 26 which is properly labeled and needs little explaining. The crushed coke and other material which are placed between the outer zinc cup and the center carbon electrode serve to hold the electrolyte with which it is soaked and to act as a depolarizer, the importance of which has already been explained. The electrolyte consists of sal-ammoniac (ammonium chloride) and zinc chloride in solution. The zinc chloride is used to counteract local action and other deteriorating effects which occur during the shelf life of the cell.

Dry cells for use as radio B and C batteries are assembled in what is known as *block* construction in which the required number of cells are connected in series and sealed in blocks of insulating material with the required terminals brought out through the

top. The cells are of small size and are arranged in a horizontal row or layer, one or more layers of cells being used depending on the voltage required.

A later development in the construction of these batteries utilizes the scheme of building up individual cells with alternate sheets of zinc and manganese-carbon mixture separated by a layer of paper with the electrolyte soaked into the paper and mixture. The individual cells are separated by sheets of conducting material which also serve to confine the electrolyte to the individual cell. The whole block is encased in the familiar paper carton.

The effect of temperature on dry cells is worthy of consideration if extensive use is made of this source of e.m.f. Ordinary temperatures have little effect on the open-circuit voltage of dry cells. Low temperatures decrease both the voltage and life of the cell. High temperatures increase local action and, therefore, decrease the life of the cell. Dry cells in storage should be kept as cool and dry as possible. The best operating temperature for dry cells seems to be about 68°F.

25. The Telephone Receiver and Transmitter.—*The function of the telephone receiver or headphone, as used in radio, is to make audible the*

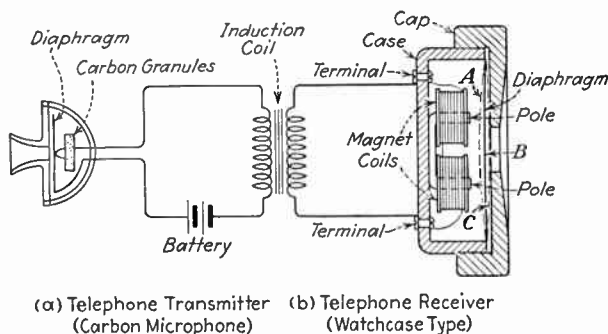


FIG. 27.—Telephone circuit showing construction and connection of receiver and transmitter.

rectified high-frequency oscillations delivered by the detector or audio amplifier. Headphones used in radio work are known as the watchcase type and are constructed as in the illustration Fig. 27 which is properly labeled. A close study of the figure (b) will show clearly the component parts of this instrument. Under normal conditions the diaphragm is in position B. When a current flows through the electromagnetic coil windings, however, the diaphragm vibrates within the limits of the dotted lines A and C, depending upon whether the current in the windings at a given instant is aiding or bucking the effects of the permanent magnets themselves. The diaphragm is like the head of a drum, and it takes only a very slight shock to make it respond loudly. Great care must be taken not to bend the diaphragm, as this, for obvious reasons, will make the headphone inoperative. The distinctive feature of the radio headphone is that it is wound with extremely fine wire, usually No. 40 B. and S. gage, which makes it possible to get a great number of turns in a

relatively small space. The m.m.f. developed by the electromagnets in the headphone depends upon the number of turns on the coil and the amount of current passing through the coil. The large number of turns results in a relatively large magnetic field being produced, even though only a feeble current is flowing. This feature makes the headphones especially responsive to weak signals. Because it is necessary to wind on so many turns of wire, the radio headphone is graded according to the resistance of its coils. The standard resistance for headphones used in commercial practice is between 2,000 and 3,000 ohms per pair. Such phones consist of two receivers connected in series each with a resistance of 1,000 or 1,500 ohms. There may be as many as 10,000 turns of wire used in making a headphone of this resistance. The resistance is always measured with direct current.

It is obvious that headphones will respond to audio-frequency currents only; and, even if the diaphragm were able to vibrate at the terrific speed of radio frequency, it would be above the hearing range of the human ear. The frequency to which the ordinary watchcase-type receiver will respond most strongly is called its *resonant frequency*.

Referring to Fig. 27, which clearly shows the component parts of a headphone of the metal-diaphragm type, it will be noted that the diaphragm is placed directly over the electromagnets. When no current is flowing through the windings of the magnets the diaphragm is in the slightly strained position *B*, due to the permanent magnetic field of the telephone magnets. However, when a current flows through the windings of the magnets, the diaphragm is drawn toward them to the position *A*, and, as soon as the current pulse has passed, it springs back to a position slightly beyond point *B*, to *C*, due to its mechanical inertia. Hence, if a permanent magnet is used it will tend to prevent the diaphragm from swinging too far beyond point *B*, and will, therefore, prevent rattling. For example, if a telephone receiver were constructed of a diaphragm and a simple electromagnet, no matter in which direction the current passed through the winding the diaphragm would be attracted. Hence, if an alternating current of 500 cycles passed through it, the diaphragm would be attracted once for every increase in current or 1,000 times a second. This would obviously result in speech distortion, for, if one should talk into a transmitter at a certain frequency, then the sound in the receiver would be of higher frequency, thus producing an unfaithful signal reproduction.

This disadvantage is greatly minimized by using a permanent magnet so that the core is magnetized in one direction constantly. Hence, the diaphragm will be continually in a bent position and thus, if a current passes through the winding in one direction, it will add to the flux (magnetic lines of force) of the permanent magnet and bend the diaphragm still more. Then when the current through the winding reverses, the flux of the permanent magnet will be weakened and the diaphragm will be released somewhat from its normal position. Thus the action of the 500-cycle current will now produce a movement of the diaphragm one way on one half cycle and the other way on the other half, and the air pulsations will then occur 500 times a second and natural-sounding speech will result.

The Carbon Microphone.—This device in the simple form used on the conventional telephone system is shown in Fig. 27. It operates by virtue

of a vibratory diaphragm which varies the resistance of a carbon-granule button. This button is in series with the line and the induction-coil primary and, therefore, governs the current flow through the circuit. When the granules are compressed, the resistance of the button is lowered and the current in the line increased and vice versa. The secondary of the induction coil is connected to the telephone receiver.

During normal operation the diaphragm of the microphone varies in accordance with the frequency of the speech vibrations (waves) impinged through the mouthpiece. These vibrations of the diaphragm in turn set up varying current in the telephone line, due to the varying resistance of the carbon-granule button, which causes a correspondingly varied e.m.f. to be induced in the secondary of the induction coil to which the receiver is connected, thereby also affecting the receiver in the same way. The frequency of the current variation in the line corresponds to the diaphragm frequency, so that a pattern of the air wave caused by the speech is reproduced in the line as an electric wave. This electric wave is again transformed into an audible air wave by the action of the telephone receiver which has already been described.

II. MAGNETISM AND ELECTROMAGNETISM

Magnetism was one of the earliest discoveries in the field of physics which later led investigators like Faraday to study the behavior of electricity. Electrical science is built on the basic principles of magnetism and magnetic effects are evident wherever electricity flows. A clear understanding of radio and electricity can be had only if magnetism and its effects are understood; it is a basic phenomenon which underlies the operation of many types of electrical and radio apparatus.

26. Natural and Artificial Magnets.—The name *magnet* was first applied to brown-colored stones which possess the peculiar property of attracting small pieces of iron or steel. Later it was found that, if a piece of such stone was suspended freely by a string, it possessed the very important property of pointing in a definite direction, very nearly north and south. It received the name *lodestone* or *leading stone* because of this directional property. In technical language this material is referred to as magnetic oxide of iron, or *magnetite*. The magnetic properties or attractive forces seem to be centered at two or more points on the stone, whereas at other points no magnetic force is evident. It is called a *natural magnet* because it is found to be magnetic in its natural state, and a lodestone or leading stone because it possesses the remarkable property referred to, which caused it to be used for navigation.

All magnets used in the construction of electrical apparatus and machinery are *artificial* magnets, that is, they must be made magnetic by some process. Hereafter in this text, when a magnet is referred to, an artificial magnet is meant.

Magnets may be divided into two classes, namely, temporary magnets and permanent magnets. The class into which a magnetic material falls depends upon its atomic structure. Most temporary magnets are made of soft iron (steel) and most permanent magnets are made of a hard-steel alloy. For practical purposes a temporary magnet is one which quickly loses its magnetism after the magnetizing force is removed, and a permanent magnet is one which retains its magnetism over a long period of time after the original magnetizing force is removed, although eventually it too may lose its magnetic property and need to be remagnetized. Both types are extensively used in radio and electrical apparatus.

If a piece of steel such as a darning needle is stroked in one direction with a piece of lodestone, it will be found to be magnetized. This can be proved by the fact that it will attract iron filings, a compass needle, or any other small piece of magnetic material. If this needle is now used to stroke a second needle, the second needle will be magnetized and so one magnet may be used to produce another magnet. This is a common laboratory demonstration, but for practical purposes this method is useless and magnets used in electrical apparatus are produced by electromagnetic means. This method is fundamentally simple and means that a magnetic material such as iron or steel is surrounded by turns of wire through which an electric current is made to flow. In this way the magnetic material is magnetized. Electromagnets will be more fully discussed later in this chapter.

27. Poles of a Magnet.—The ends of a magnet are termed its *poles*. The end which points to the north geographical pole is generally called the *north-seeking pole*, and is usually marked on one end of the magnet by an N while the other end is called the *south-seeking pole* and is marked S.

The term *magnetic polarity* refers to the nature of the magnetism at a particular point on the magnet, or in other words, whether it is a north- or a south-seeking magnetism.

Magnetic Attraction and Repulsion.—If

a steel bar which has been magnetized is suspended at the center by a silk thread and placed in the vicinity of a north or a south magnetic pole, it will tend to swing parallel with the magnetic field created by the pole.

Figure 28 illustrates the magnetic lines of force emanating from the *north* pole and entering at the *south* pole through the air medium. The lines of force then return to the *north* through the magnet, thus completing its field.

If the bar is carefully marked at one end for reference, it will be noted that one particular end will always point in a definite direction. This is

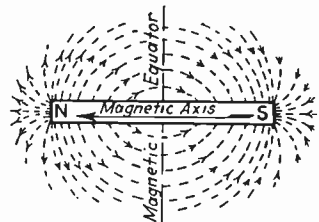


FIG. 28.—Poles of a magnet, and magnetic field.

due to the polarity effect of one magnetic field upon another. No matter how many times the bar is turned by hand it will always return to its original magnetic position. Figure 29(a) and (b) illustrate the magnetic effect of two magnetic fields, one upon the other. Figure 29(a) shows how the lines of force tend to combine and thus attract the two bodies, while Fig. 29(b) shows the "bucking" or repelling effect of the fields and, hence, a repulsion of the two bodies. This holds true with all

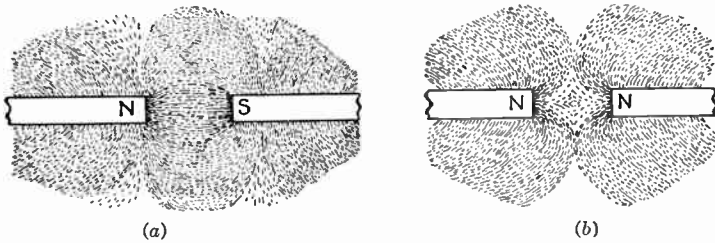


FIG. 29.—Magnetic effect between; (a) unlike, and (b) like, poles of a magnet.

magnetic bodies. Whenever two *north* poles or two *south* poles are placed near each other, they will tend to repel one another. On the other hand, if poles of *north* and *south* polarity are placed in the same vicinity they will attract each other owing to the combining effect of the magnetic fields.

If the strength or distance between the magnets is varied the effect will vary accordingly.

This gives us the laws of magnetism which are:

1. *Unlike poles attract one another: like poles repel one another.*
2. *The force between two magnetic poles varies inversely as the square of the distance between them and is proportional to the product of their strengths.*

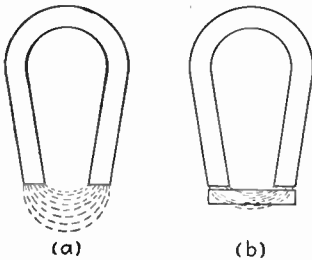


FIG. 30.—Magnetic field arrangement of a horseshoe magnet.

The second law is not strictly true because it assumes each pole is a point or dot. It is true enough for practical purposes, however, provided the poles of the magnet are not too close to one another and are long in proportion to their cross-sectional areas.

28. Magnetic Fields.—Various forms of magnetic bodies possess certain magnetic advantages. Of these forms the two most important are the bar magnet and the horseshoe magnet.

Figure 30(a) illustrates the magnetic-field arrangement of a horseshoe magnet. Compare this with the field arrangement of the bar magnet in Fig. 28. Note that the concentration of the lines of force in Fig. 30(a) is much greater than in Fig. 28. This is obviously due to the shorter air path through which the lines of force travel, and hence results in a lower

amount of magnetic leakage. Therefore the strength of the magnetic field at the poles of a horseshoe magnet would be considerably stronger than that of a bar magnet. If it is desired, however, to decrease the magnetic leakage of the horseshoe to a still lower degree, then a small iron bar may be placed across or between the poles as illustrated in Fig. 30(b).

Whenever a low-loss magnetic circuit is required, the magnetic material may be formed into a closed magnetic path by arranging it into a horseshoe, circular, or rectangular form with a small air gap between the poles.

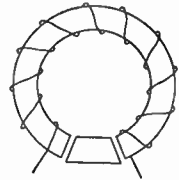


FIG. 31.—The ring magnet.

The ring magnet (Fig. 31) provides a complete metallic path for the magnetic field. As the magnetic path is continuous, this magnet has no poles. If the ring is broken at any point a north and south pole will immediately be produced at the ends. A transformer core is an example of a ring-type magnet in rectangular form. The ring magnet is also used in certain electrical indicating instruments.

Characteristics of Flux Lines.—Flux lines have certain characteristic tendencies which may be stated as follows: they are continuous throughout the magnetic circuit and form closed lines; they act like stretched rubber bands, tending to shorten themselves while at the same time a stress exists which tends to keep each line separate from its neighbor; the magnetic field at any point is the resultant of all of the lines acting on that point.

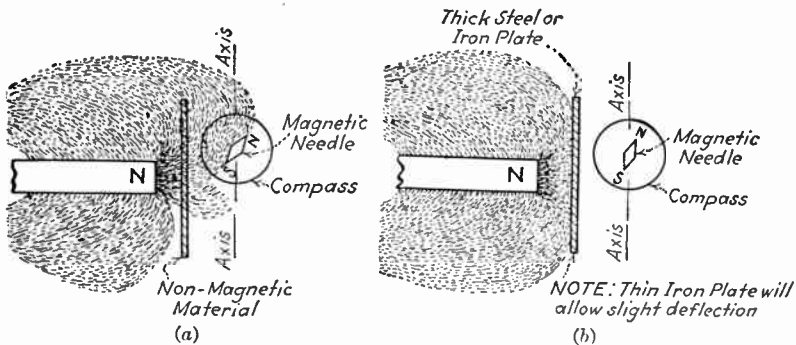


FIG. 32.—(a) Non-screening effect of non-magnetic material; (b) screening effect of magnetic material.

29. Magnetic Transparency and Leakage.—If a magnet were placed in a position to deflect a magnetic needle, and if a non-magnetic substance, such as glass, wood, brass, or rubber, were placed between the needle and the magnet, the lines of force would complete their circuit through the non-magnetic body, as shown in Fig. 32(a). If a piece of iron were interposed between the magnet and the needle, however, the iron would

act as a *magnetic screen* and reduce the needle deflection toward the magnet (Fig. 32(b)).

It is seen then that a magnetic material may be used to screen off a magnetic field. This screening effect is due to the lower reluctance of the magnetic screening material which causes the magnetic lines to take the easiest path through the magnetic screen.

In electrical machines and instruments the magnetic field is designed to act on certain windings and so produce a specific effect. Only those lines of force or that flux which so acts as intended may be called *effective flux*. A certain amount of flux is always lost in the surrounding space, although this loss may be kept small if desired. The flux which is lost and is therefore ineffective is said to be lost through *magnetic leakage*. Magnetic leakage may be desirable under some conditions and undesirable under others.

30. Theory of Magnetism.—Although magnetism has been studied for over a century, no satisfactory theory has yet been evolved to explain all magnetic phenomena. There is, however, a mass of disconnected facts, many of which lend themselves readily to the evolution of a theory of magnetism.

Modern theories of magnetism, such as they are, are primarily due to the basic work on magnetism done by Professor P. Curie of radium fame. It is logical that Curie's work on magnetism and radium should be so closely interrelated, as modern science now recognizes the relationship of the two fields of research.

Recent research has uncovered the fact that the atom, in addition to being made up of electrons and protons, contains a third component part called *electromagnetic energy* which it gives off in the form of radiation under certain conditions. It is this released energy which is now thought to be the cause of magnetic effects. This belief has grown out of the modern theory of magnetism which attributes magnetic phenomena to the rotation of electrons inside the atoms of magnetic substances.

The electron theory of magnetism, which seems to be the most acceptable of all magnetic theories thus far advanced, depends on the existence of magnetons, which in

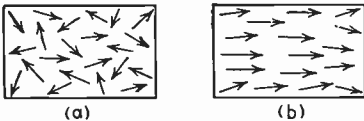


FIG. 33.—Polar arrangement of atoms in a molecule. (a) Unmagnetized material; (b) magnetized material.

magnetic terminology and structure are equivalent to the electron in electrical terminology and structure. Just as an atom contains electrons so does it contain magnetons, so many to each atom, depending on the material. These magnetons are, in effect, magnetic forces which make of each atom an elementary magnet.

In an unmagnetized material it is supposed that the magnetic forces, the elementary magnets within a molecule, are directed in all directions in a promiscuous fashion as indicated in Fig. 33(a). When magnetism is induced into the material, it is supposed that each molecule becomes polarized owing to a reorientation of the

atoms within the molecule, so that the magnetic forces within all of the molecules are directed in the same direction [see Fig. 33(b)].

Weiss, a noted investigator of magnetism, claims an extra or intrinsic field in addition to the individual atomic magnets, and that to this molecular field must be attributed the action of forces, neither magnetic nor electrostatic, whose nature is still unknown. Perhaps this unknown force may prove to be that component of the atom now known as *electromagnetic radiation*.

If a material is placed in a magnetic field in air, the lines of force will either be closer together or farther apart as they pass through the material. If the lines are farther apart, the material is known as a *diamagnetic* material or, if closer together, as a *paramagnetic* material. Iron, and materials of which iron is a divalent, nickel, and cobalt are classed together and are called *ferromagnetic*. Ferromagnetic materials are, therefore, a kind of paramagnetic material. Owing to the action of the lines of force on the magnetic material, a magnet will repel slightly a diamagnetic material and attract slightly a paramagnetic material, unless it is a ferromagnetic material in which case it will attract it strongly.

Tin, zinc, silver, copper, gold, and lead are a few of the diamagnetic materials. All iron alloys, nickel, and cobalt are ferromagnetic, and all other magnetic materials are simply classed as paramagnetic. This sums up to the fact that all materials less magnetizable than air are called diamagnetic; and all materials more magnetizable than air are called paramagnetic. Nearly all magnetic material with which the electrical or radio engineer has to deal are of the ferromagnetic class.

Langevin, another authority on magnetic research, states in effect that substances whose atoms have their electrons in revolution in such a way that their effects are additive are paramagnetic and that the atoms of such substances may be looked upon as elementary magnets.

The spacing of the atoms within the molecule seems to play a very important part in magnetic phenomena. This is evidenced by the effect of heat on a magnetic material. Heat increases the space between the atoms and eliminates magnetism.

*Magnetostrictive effects*¹ cannot as yet be explained by the theories thus far outlined. Simple orientation of the molecules would not explain a change in the physical dimensions of a ferromagnetic bar unless it is assumed that the molecules take up more room when oriented a certain way. But this theory seems to fall down when it is considered that an increase in magnetizing force may cause the material both to expand and to contract.

Magnetic theory in its present status is far from settled, and new ideas are continually being advanced. Some day, science may evolve a satisfactory theory to explain *all* magnetic effects but until then the pres-

¹ See *Magnetostriction Effect*, Sec. 33.

ent hypothesis called the *electronic theory of magnetism* will serve for most purposes.

31. Magnetic Variables.—A magnet may lose its magnetism if subjected to a physical shock. The amount of magnetism lost will depend upon the severity of the jar. A new magnet will lose a greater percentage of magnetism than an old magnet, if both are subjected to the same shock. Any piece of apparatus, such as a meter or telephone receiver, containing a permanent magnet should be carefully guarded against rough usage.

A magnet will lose its magnetism upon being heated to a bright-red heat. Lesser heating will cause a magnet to lose a proportionate amount of its magnetism. If a permanent steel magnet is placed in hot or boiling water its magnetic strength will be reduced, although it will be partly recovered when the magnet cools off. Chilling a magnet increases its magnetic strength.

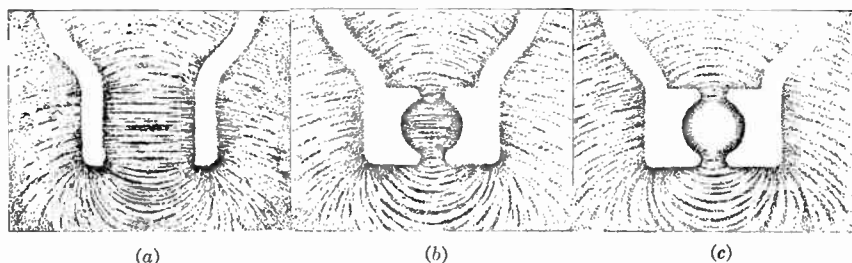


FIG. 34.—Magnetic field between the poles of a magnet. (a) Normal field; (b) concentrated field; (c) radial field due to armature. (Courtesy of Weston Electrical Instrument Corporation).

Constancy of magnetic strength in a magnet is by no means permanent. When a permanent magnet is first magnetized—when it is being made—it is found that the magnetism consists of stable and unstable magnetism. The unstable magnetism must be removed, before the magnet is put to practical use. Even the stable magnetism decreases with the age of the apparatus. Every physical shock, every contact the magnet might make with iron, every change in temperature, all of these factors tend to weaken its magnetism until eventually its magnetism becomes dissipated below a useful degree and the magnet must be remagnetized. Remagnetization of the magnets in a direct-current generator is sometimes necessary to replace residual magnetism. It is frequently necessary to remagnetize the permanent magnets in radio telephone receivers, especially the lower resistance types.

32. Purpose of Pole Pieces.—Only those lines of force from a permanent magnet which traverse the windings of a meter or rotating machine are useful lines of force. Normally the field between the poles of a magnet consists of straight lines of force as shown in Fig. 34(a). A concentration of these lines of force is desired so that the field will be confined

where it will be most effective, and a more uniform field maintained throughout the armature windings. This field concentration may be accomplished by fitting pole pieces to the poles of the magnet resulting in a field as shown in Fig. 34(b). The ideal position of the lines of force between the pole pieces is radial, like the spokes of a wheel. With this condition the lines of force cut the armature windings at a uniform angle throughout their entire revolution. The effect of the field on the windings is then more uniform, and better operating results are obtained from a unit section of the field. This is accomplished when the armature core is placed between the pole faces. For a two-pole meter, the field is as shown in Fig. 34(c). For a multi-pole generator, the field is as shown in Fig. 35. Note that the lines of force are parallel to the radius of the core and are therefore radial.

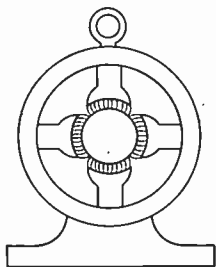


FIG. 35.—Radial field produced in generator by armature.

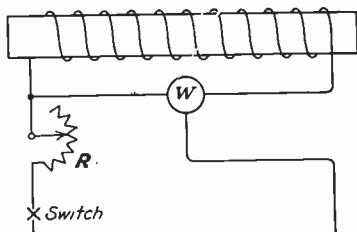


FIG. 36.—Diagram to illustrate magnetostriction effect.

It is to be noted also that the curved faces of the pole pieces, fitting closely around the armature, reduce the air gaps between the pole faces and the armature to a minimum, thereby reducing stray magnetic losses to a minimum.

33. Magnetostriction Effect.—To illustrate the magnetostriction effect, suppose first an unmagnetized steel bar is inserted within a solenoid connected to a source of e.m.f., the resistance R is at a maximum value, and the switch is open (see diagram, Fig. 36).

If the switch is now closed, a reading will be observed on the wattmeter indicating power in the electromagnet. If a direct-reading micrometer-indicating instrument were attached to measure the length of the bar, it would be found that when the bar was magnetized by the field from the solenoid it would contract a few millionths of an inch. If the value of the resistance is now decreased, an increased power will be noted on the wattmeter. The bar will now return to its original length owing to the increased magnetic effect. The bar has thus far completed one alternation of physical vibration.

If the value of R is still further decreased, allowing more power into the electromagnet, the bar will contract its length. This contraction and expansion continue as long as the state of magnetism in the bar is varied, and at a rate depending upon the physical dimensions of the bar and the magnetic material used.

The conclusion drawn from this experiment is that a change in magnetic condition causes a change in physical condition in the bar. This is called the *direct* magnetostriction effect.

Another magnetostriction effect has been observed which is the reverse of the direct effect just described. This effect is that a bar of magnetic material whenever it is subjected to a mechanical stress will change its magnetic state. This effect is generally called *indirect* magnetostriction.

Now, if the steady e.m.f. applied to the electromagnet is replaced with an alternating e.m.f., the frequency of which is resonant with the vibration period of the rod, the distorting effect on the rod may be increased 100 times.

34. Fields around Conductors.—The normal position of the electrons and protons in a wire before the application of a potential may be visual-

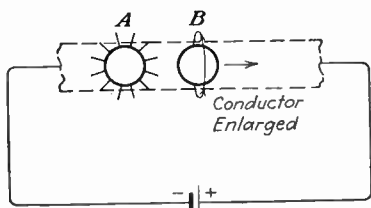


FIG. 37.—Movement of an electron in a wire.

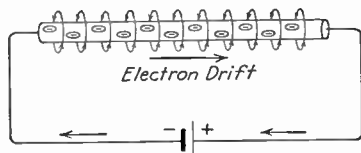


FIG. 38.—Electronic drift in a conductor.

ized as a number of positive and negative charges with their electrical forces acting upon one another. The negative forces about the electrons and the positive forces about the protons being equal, there will be no exterior effect on the outside of the wire. Even the movements of the free electrons in conducting materials produce no exterior effects on the outside of a conductor because they move relative to each other in all directions with equal velocity and consequently their average vector velocity is zero.

If, however, a difference of potential is applied to the conductor, the free electrons in the conductor will be displaced in a certain direction, depending upon the polarity of the applied e.m.f., which results in a progressive drift of the electrons along the conductor.

Now, since the electrons are known to have an electric field about them extending in straight lines and in all directions, the progressive motion of the electrons along the conductor will cause a distortion of their electric lines in such a manner as to produce an exterior manifestation at right angles to their progressive motion. This condition is illustrated in Fig. 37 in which an electron is displaced from left to right. Thus, if a number of free electrons in a conductor are moving progressively from

left to right, an exterior manifestation of the displaced electric fields will result. This external field is called the *magnetic field* and is the theoretical basis of operation of all dynamo-electrical machinery.

Figure 38 illustrates the magnetic field around a conductor carrying current. It will be seen that the electrons are moving toward a point of high potential (+) and that their motion results in a magnetic field in a given direction. It is important to note at this point that the direction of the exterior magnetic field is entirely dependent upon the progressive motion and the direction of the electrons in the conductor. For example, if the electrons are moving along a conductor away from the observer, the magnetic field will be in a counterclockwise direction, and, if the electrons are moving toward the observer, the direction is clockwise. This condition is illustrated in Fig 39 and may be permanently remembered by using the left hand as a rule. The thumb points in the direction in which the electrons are flowing (toward the positive potential), and the fingers indicate the direction of the magnetic lines of force (see Fig. 42).

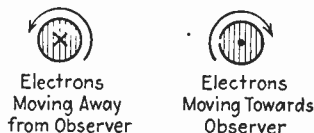


FIG. 39.—Magnetic field around a conductor.

In all practical circuits the magnetic field is looked upon as a separate entity, but in reality it is always associated with an electric field. It is therefore convenient to look upon a magnetic field around current-carrying conductors as the manifestation of forces (magnetic) evoked by the movement of an electric field (electrons in motion).

Since each electron in motion contributes a proportionate share of the current flow in the conductor, it can readily be seen that the greater the number of electrons which are in progressive motion, the larger will be the current flow, or amperes, in the conducting circuit. Furthermore, since each electron in progressive motion also contributes its external magnetic forces (magnetic field), the greater the number of electrons moving along the conductor, the stronger will be the magnetic field. Each electron in motion may be pictured as possessing a magnetic line of force and, since there are a great number of electrons in motion through a conductor, the exterior magnetic field will possess more lines of force or a stronger magnetic field.

The presence of a magnetic field about a conductor carrying current may readily be determined by placing a magnetic needle or compass in

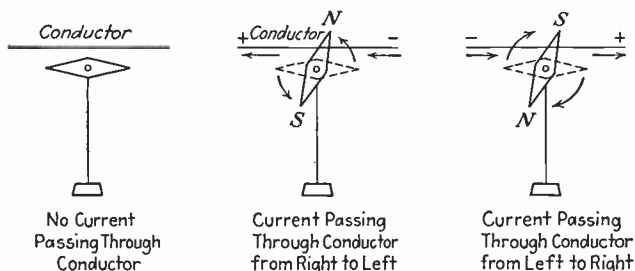


FIG. 40.—Effect of magnetic field on a compass needle.

the vicinity of the conductor, as illustrated in Fig. 40. When the circuit is closed and the electrons are caused to flow in the direction of the arrows, the needle will tend to turn at right angles to the conductor; but, if the flow is turned off, the needle will resume its original position. The

force which made this needle turn at right angles to the conductor was the lines of force, or magnetic field, produced by the progressively moving electrons in the conductor. The magnetic field is definitely proved to be in this right-angular position because it was in this direction that the needle turned when placed in the vicinity of the field.

Now, since the direction of a magnetic field about a conductor carrying current is definitely known, it is possible to determine the effect of these fields upon one another in accordance with the fundamental laws of



FIG. 41.—Effect of (a) supplementary and (b) opposing fields in adjacent conductors.

attraction and repulsion of magnetic bodies in the vicinity of one another. For example, if two conductors are placed side by side and are carrying current in the same direction, the magnetic forces will also be in the same direction, or supplementary, which will cause the combining of the fields into one larger and consequently stronger magnetic field as shown in Fig. 41(a). If the current flow in adjacent conductors, on the other hand, is in opposite directions, then the magnetic forces will oppose or neutralize each other and may actually result in the complete cancellation of the exterior effectiveness of the magnetic fields, as shown in Fig. 41(b).

The strength of a magnetic field around a conductor may be greatly increased by forming the wire into a coil or solenoid, thereby causing the lines of force around each wire to interlink or combine with the adjacent or succeeding turns.

The magnetic field through a coil of this kind is similar to that of an ordinary bar magnet in that the lines of force all concentrate in the center or axis of the coil and *leave* and *enter* at the magnetic poles. The strength of the magnetic field around a coil may be greatly increased by inserting an iron core or bar of soft iron therein.

Thus it can be readily seen that the strength of a magnetic field around a coil or solenoid may be increased by one or more of the following methods:

1. Increasing the electron flow (current) through a conductor.
2. Increasing the number of turns on the coil.
3. Increasing the diameter of the coil.
4. Increasing the flux density by inserting an iron core.

*Left-hand Rules.*¹—Several simple rules have been devised to aid in remembering the relationship between the direction of the magnetic lines

¹ Based on the fact that current flows from negative to positive.

of force and the electron flow. Figure 42 illustrates an easy way to remember and determine the direction of the magnetic field around a conductor if the direction of the electron (current) flow is known. The conductor is grasped as shown, the thumb pointing in the direction of the electron flow. The fingertips then point in the direction taken by the lines of force as they build up and surround the conductor.

The magnetic polarity of a coil or solenoid may be determined in a similar way as shown in Fig. 43. With the left hand, the coil is grasped so that the fingers point in the direction of the electron flow. The thumb will then point in the direction of the north pole of the magnetic field.

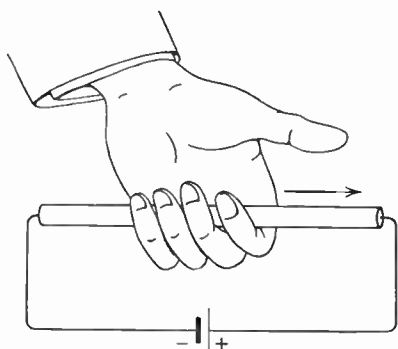


FIG. 42.—Finding the direction of the magnetic field in a conductor by the left-hand rule.

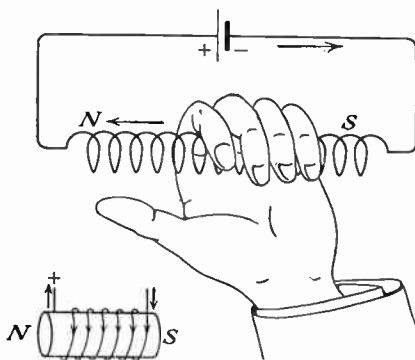


FIG. 43.—Finding the direction of the magnetic field in a solenoid by the left-hand rule.

35. Induced Currents.—Faraday discovered that, if a closed-circuit conductor such as a coil was moved in the vicinity of a magnet, a current would flow in the circuit. He followed this up later by the discovery that a varying current in one conductor would cause a current to flow in a second conductor, provided the second conductor was brought close enough to the first one and a continuous path provided for the current to flow in. Such currents are said to be generated by *induction* and are called *induced currents*. The combined action of induction and current flow is termed *electromagnetic induction*.

Electromagnetic induction is thus responsible for the operation of all electric generators and motors, and, when this phenomenon is used for this purpose, it is termed the *dynamo-electrical effect*. Electromagnetic induction is also responsible for the transference of electrical energy from one closed circuit to another and when so used is termed the *transformer effect*.

Law of Induction: 1. When an e.m.f. is induced in a closed circuit by a conductor cutting a field, or vice versa, the amount of current flow is proportional to the rate of cutting and the number of linkages.

2. The induced e.m.f. sets up a current the direction of which tends to prevent a change in the number of linkages.

The above statements, which are proved to be true, have been summarized by Lenz into a law known by his name which substantially concludes as follows: *An induced e.m.f. always sets up a magnetic field the direction of which is such as to oppose any change in the force (field) which produces it.*

36. Production of E.M.F. by Field Cutting Conductor.—A current flow can be set up in a conductor by moving it through a magnetic field.

Figure 44 illustrates this fundamental action. A permanent magnet is placed so that a conductor can be moved in front of its pole faces. A sensitive zero-center galvanometer is connected to the ends of the conductor to indicate the e.m.f. induced. If the conductor is started above the north pole in the position *A*, shown by the dotted lines, and swung down in front of the north pole, a deflection will be noted on

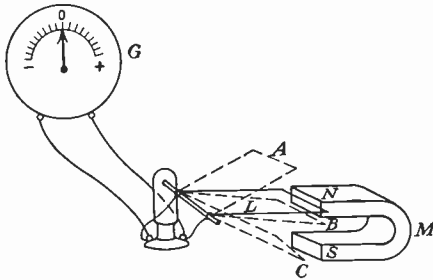


FIG. 44.—Experimental set up for the production of an e.m.f. by a field cutting a conductor.

the galvanometer. This deflection will be, let us say, for the sake of illustration toward the positive sign on the scale. The value of e.m.f., as shown by the meter, will be a minimum at the start of the movement and maximum as the conductor passes directly in front of the pole face, returning to minimum at position *B* which is the point where the conductor passes from the field coming out of the north pole and enters the field going into the south pole.

As the direction of the field in relation to the conductor is reversed as the conductor continues its journey, and begins to pass the south pole, the induced e.m.f. will be reversed and will be so indicated on the galvanometer which will swing over toward the negative sign on the scale. The value of induced e.m.f. will follow the same general pattern of rising from zero to maximum and back to zero again as it passes in front of the south pole just as it did when passing the north pole, the only difference being in the reversal of the induced e.m.f.

If the conductor is swung upward from position *C* to position *A*, currents will be induced in the conductor, and they will be in the same order to what they were when the conductor was swung downward as before explained. The galvanometer would indicate this condition by swinging first toward the positive and then toward the negative as the conductor passed first the south pole and then the north pole.

If the conductor is moved slowly across the front of the magnet, the deflection on the meter will be relatively small, whereas, if the conductor is moved quickly through the magnetic field, the deflection on the meter will increase.

If the flux in the magnetic field is increased by utilizing a winding around the magnet and making it an electromagnet, the deflection on the meter will be greatly

increased. The increase in induced e.m.f. will be proportional to the increase in the field flux.

Another way to increase the induced e.m.f. would be to utilize more than one turn of conductor. Two turns will cause twice the deflection on the meter that one turn will cause, everything else being equal. It is apparent, then, that the more turns, the greater the induced e.m.f.

From this experiment we are able to deduce certain fundamental laws governing the generation of a current by the dynamo-electrical effect, namely: (1) whenever there is a relative movement between a conductor and a magnetic field, an e.m.f. is induced in the conductor; (2) the polarity of the e.m.f. induced in the conductor depends upon the direction of the relative motion between the conductor and the magnetic field; (3) the intensity of the e.m.f. produced in the conductor depends upon the intensity of the magnetic field and the speed of the relative motion between the conductor and the field; (4) the value of the induced e.m.f. varies directly with the number of turns in the conductor.

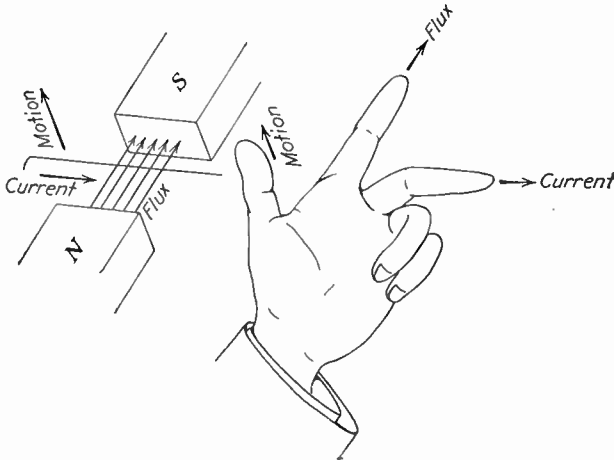


FIG. 45.—The left-hand generator rule (according to the electron theory).

Left-hand Generator Rule.—A rule which will give the relation between direction of flux, motion, and direction of electron flow is stated below. *Point the forefinger of the left hand in the direction of the flux, the thumb in the direction of motion; the middle finger will point in the direction of the electron flow in the metallic circuit.* This rule is illustrated in Fig. 45.

III. THE MAGNETIC CIRCUIT

37. The magnetic circuit is similar to the electric circuit in that in it is provided a force of a unit quantity which overcomes a unit resistance. In the electric circuit these factors are measures of e.m.f., current, and resistance, whereas in the magnetic circuit they are represented by

m.m.f., flux density, and reluctance. An Ohm's law of the magnetic circuit may, therefore, be drawn up as follows:

$$\text{Flux} = \frac{\text{m.m.f.}}{\text{reluctance}}$$

a. *Flux* is the total number of lines of force making up the magnetic field. The symbol for flux is ϕ , pronounced phi.

b. *Magnetomotive force* (m.m.f.) is comparable to electromotive force in that it is a force producing a certain phenomena. It is represented by the letter F . The unit of measure of m.m.f. is the *gilbert*, and 1 gilbert is equal to 1.26 times the number of turns times the current flowing in the circuit or,

$$F = 1.26 \times N \times I \quad (23)$$

c. *Reluctance* is that property of a material which opposes the creation of a magnetic flux in the material. It is the "resistance" of the magnetic circuit or the counter force in the magnetic circuit which the magnetic field must overcome. The symbol for reluctance is the script letter \mathcal{R} , and the most commonly accepted unit of reluctance is the *oersted* which is the reluctance of 1 cc. of air.

Reluctance may also be expressed in *rels*. A material has a reluctance of 1 rel when an m.m.f. of 1 ampere-turn generates in it a flux of one line of force. Ampere-turns = $N \times I$, where N is the number of turns and I is the current in amperes.

d. *Reluctivity* is a measure of *specific reluctance*, which may be expressed in oersteds per cubic centimeter or rels per cubic inch. If the c.g.s. system is used, the reluctivity of all materials including air, but excepting iron and steel and their alloys, is 1 oersted. If measurement is made in cubic inches, 0.313 rel is used as the specific reluctivity figure for air, instead of 1 oersted. The symbol for reluctivity is ν , the Greek letter nu.

Reluctance, like resistance, varies inversely with the area, and directly with the length, of the path. For a given magnetic circuit the reluctance is

$$\mathcal{R} = \frac{\nu l}{A} \quad (24)$$

where ν is specific reluctance in oersteds or rels, l is the length of the magnetic circuit in centimeters or inches, and A is the area or cross section of the magnetic path in centimeters or inches. If ν is in oersteds, l and A must be in centimeters; and if ν is in rels, l and A must be in inches.

e. *Magnetic and Non-magnetic Materials*.—Although all material is magnetic to a certain extent, it is customary to refer to all matter including air, but excepting iron and steel and its alloys, as non-magnetic. Iron and steel and its alloys are referred to as magnetic materials.

38. Ohm's law for magnetic circuits is summarized as follows:

$$\phi = \frac{F}{\mathcal{R}} \quad (25)$$

$$\mathcal{R} = \frac{F}{\phi} \quad (26)$$

$$F = \phi\mathcal{R} \quad (27)$$

where ϕ = flux, F = m.m.f. in gilberts, and \mathcal{R} = reluctance in oersteds. If F is expressed in ampere-turns, \mathcal{R} is usually in rels. Ampere-turns may be changed to gilberts by multiplying by 1.26 which then gives \mathcal{R} in oersteds.

39. Magnetic-circuit Computations. *a. The Gauss.*—In order to reduce the labor involved in making magnetic-circuit computations it was found necessary to provide tables and curves of magnetic characteristics of certain materials. To make these curves and tables practical it was necessary to express flux and m.m.f. in definite units. Flux, therefore, in a magnetic material is referred to as *flux density* and is expressed in *gausses per square centimeter* or *lines per square inch*. A gauss may be defined as *one line of force*.

For practical purposes the expression gauss means lines of force per square centimeter. If gauss is given in square centimeters, the m.m.f. necessary to produce this flux is given in gilberts per centimeter. If flux density is given in lines per square inch, then m.m.f. is given in ampere-turns per inch.

The symbol for flux density is \mathcal{B} , gauss per square centimeter (or B, meaning lines per square inch).

Another value which must be used in magnetic-circuit calculations is that of the m.m.f. involved. This factor is expressed in gilberts per centimeter necessary to produce a given flux density. The symbol for gilberts per centimeter is the script \mathcal{H} , or H ampere-turns.

b. \mathcal{B} - \mathcal{H} Curves.—A magnetization curve may be drawn showing the relationship between m.m.f. and flux density for a specific unit of material. Such curves are known as B-H curves. A set of these curves for various common magnetic materials is shown in Fig. 46. A typical curve for permalloy, which is commonly used in high-grade audio transformers, is shown in Fig. 47.

As the designing engineer is interested in practical units, H is usually given in ampere-turns. If \mathcal{H} is given as gilberts, the answer in ampere-turns may be secured by dividing by 1.26. B-H curves are very useful in figuring magnetic computations.

Example: Suppose a transformer is to be designed having an electrical sheet-steel core with a cross-sectional area of 4 sq. cm. It is necessary to produce a flux of 60,000 lines in the core. The length of the magnetic path is 28 cm. How many ampere turns are required to produce this flux?

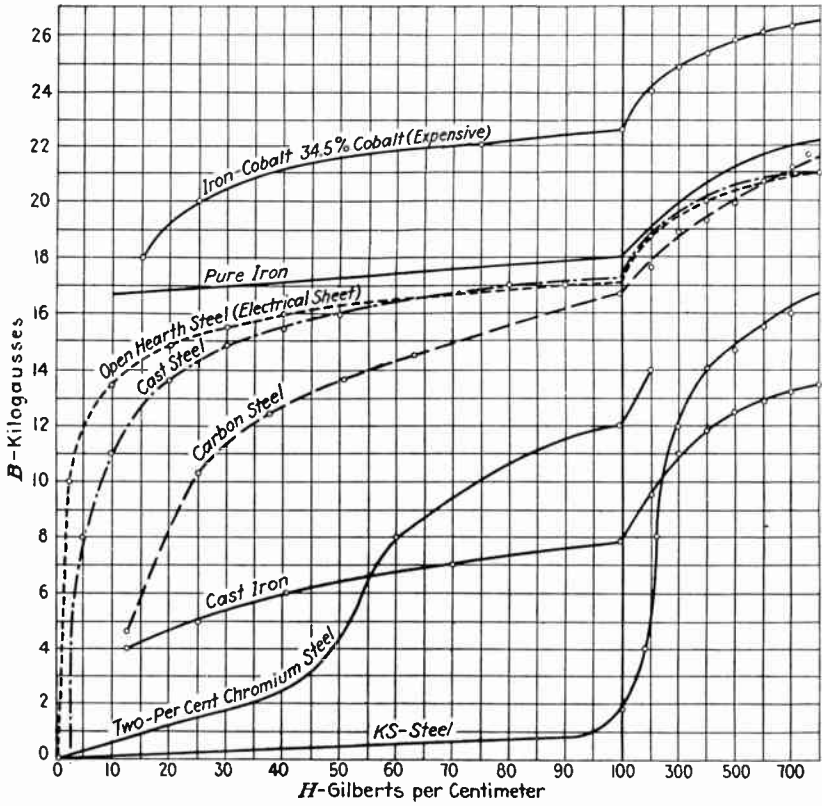


FIG. 46.—Typical B-H curves.

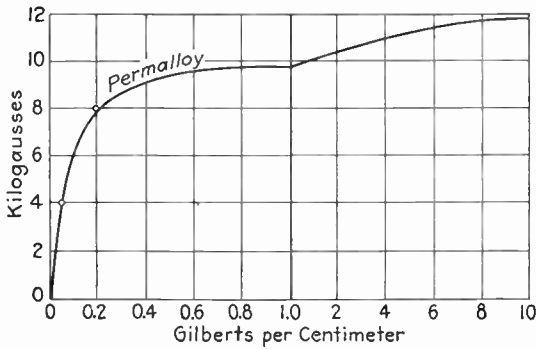


FIG. 47.—B-H curve for permalloy.

Formula:

$$\mathfrak{B} = \frac{\phi}{A} \quad (28)$$

Solution:

$$\mathfrak{B} = 60,000/4 = 15,000 \text{ gausses} = 15 \text{ kilogausses}$$

By referring to the curve (Fig. 46) for electrical sheet steel, it is found that to magnetize this material to a flux density of 15 kilogausses requires 22 gilberts per centimeter. For this path of 28 cm., 22 must be multiplied by 28 giving 616 gilberts required. Dividing by 1.26 gives 489 ampere-turns required.

c. Permeability is the comparable unit of measurement showing the ease with which a magnetic field may be set up in a material as compared with air. It may be expressed in the formula,

$$\mu = \frac{\mathfrak{B}}{\mathfrak{C}} \quad (29)$$

where μ is the permeability, \mathfrak{B} is the flux density expressed in gausses or lines per square inch, and \mathfrak{C} is the m.m.f. in gilberts per centimeter, or ampere-turns per inch. Therefore,

$$\mathfrak{C} = \frac{\mathfrak{B}}{\mu} \quad (30)$$

and

$$\mathfrak{B} = \mathfrak{C}\mu \quad (31)$$

d. Saturation.—When the molecular rearrangement of the magnetic material has been completed by the magnetizing force, a material is said to be saturated. Flux density may be increased beyond the saturation point, but the m.m.f. necessary to do this will be much greater proportionately after saturation than before saturation, and the circuit is not efficient. Inasmuch as this is true, it is practically assumed that the saturation point marks the limit of magnetization.

e. Reluctance of Magnetic Material.—While the reluctance of non-magnetic material is constant at 1 oersted per cubic centimeter or 0.313 rel per cubic inch, the reluctance of magnetic materials varies over a wide range dependent upon the material and the m.m.f. involved. For example, while the reluctance of gray cast iron is such that with an m.m.f. of 100 gilberts a flux density of 8,000 gausses may be produced, at 200 gilberts the reluctance has changed so that 9,600 gausses is produced. Using certain kinds of wrought-iron forgings, 100 gilberts will produce 16,500 gausses; whereas 150 gilberts will produce almost 18,000 gausses. To ascertain the relative reluctance of a certain magnetic material, reference may be made to permeability or induction tables or curves found in electrical engineering handbooks.

f. Remanence is a measure of the magnetism left in a material after the magnetizing force has been entirely removed and the material left in

a free state. This remainder is also called the *residual magnetism*. The property of a magnetic material which makes it possible to acquire residual magnetism is called *retentivity*.

40. **Hysteresis** is a lagging of magnetizing effect behind magnetizing cause, a lagging of \mathcal{B} behind \mathcal{H} . This may be understood by referring to the hysteresis loop or magnetic cycle shown in Fig. 48. Imagine a piece of iron which has not been magnetized and which is now to be magnetized. The curve *a-b* shows the magnetization characteristic as the m.m.f. is

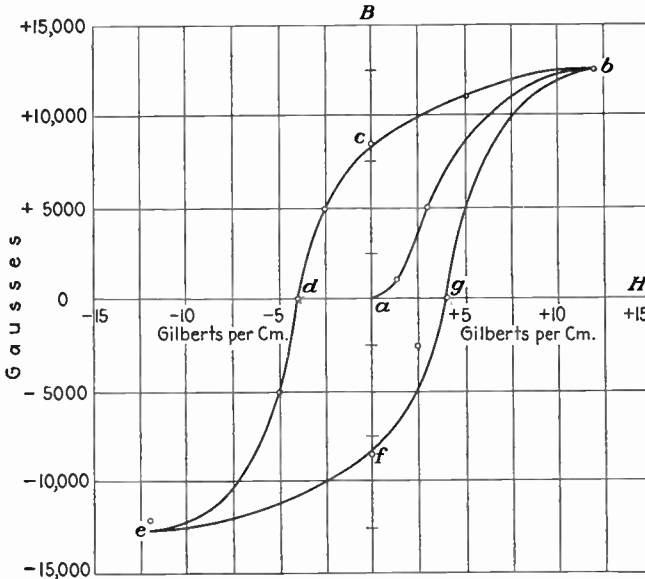


FIG. 48.—A typical hysteresis loop.

applied. If, when point *b* is reached, the magnetizing current is reduced, the iron will commence to become demagnetized. The flux density in the iron will become less as the m.m.f. is reduced until point *c* is reached where the applied m.m.f. is zero. It is to be noted that the flux density has not dropped to zero with the m.m.f. As a matter of fact a considerable amount of magnetism, or remanence, is still present in the iron owing to the retentivity of the iron. This residual magnetism, therefore, is represented by the ordinate *a-c*. As the cycle continues, a negative m.m.f. is applied; but 4 gilberts of negative m.m.f. must be applied before the iron is demagnetized. This power, which is necessary to nullify the residual magnetism, is called the *coercive force* and is a distinct loss of power. It is a hysteresis loss and may be figured and allowed for as such.

If the negative m.m.f. is increased beyond *d* to a negative value numerically equal to its positive value, the flux density will increase to the point *e* which will equal the density at *b*. If the m.m.f. is now

decreased, the same action will be repeated from point *e* to point *f* as took place from point *b* to *c*. As the m.m.f. is now applied in a positive direction the same action takes place, but in the opposite direction, from the point *f* to point *b* as from the point *c* to point *e*. The hysteresis loop or magnetic cycle is now completed. The energy lost through hysteresis is proportional to the area of the hysteresis loop and manifests itself as heat.

CHAPTER II

ALTERNATING-CURRENT ELECTRICITY

Although most electrical currents are in themselves movements of electrons along a wire, the characteristics and effects of these currents become complex if the direction or magnitude of these currents is changed in any way. Usually changes in electrical circuit constants and characteristics are due to the effects of alternating current, and it is the purpose of this chapter to explain these effects and their causes.

In alternating-current circuits the electrons do not actually traverse a circuit progressively but oscillate to and fro owing to the constantly changing polarity of the applied e.m.f.

I. ALTERNATING-CURRENT GENERATION AND TRANSFORMATION

41. The Alternator Principle.—Generators of alternating current are usually known as *alternators* and are theoretically the simplest type of dynamo-electrical machines. Basically an alternator consists of an electromagnet and a moving conductor, an e.m.f. being induced in the conductor by its movement through the magnetic field.

A simple illustration of the alternator principle is shown in Fig. 49, in which a conducting loop *ABCD* is arranged in a position so that it can be rotated on its horizontal axis in the air space between the two poles of a magnet, N and S. The loop has an outlet to two slip rings and brushes from which the alternating e.m.f. and current can be drawn for external use.

It has already been shown in the preceding chapter how an e.m.f. may be induced in a conductor by causing it to cut a magnetic field. With this theory in mind we may proceed to the analysis of the generation of an alternating current by an alternator.

Position I (Fig. 49) is the starting point of the rotation movement of the loop, and, incidentally, represents the neutral position of the loop with respect to the field from N to S. At this instant, the sides of the loop are moving in a direction parallel to the lines of magnetic flux, and the induced e.m.f. is zero as shown in the curve.

Position II shows the loop after it has revolved through an angle of 90 degrees during which time it has cut a certain number of magnetic lines of force and has induced in it an e.m.f. which in turn creates a current flow in the conductor in the direction *ABCD*. This current flow is regarded as positive in the graphical illustration and is indicated by the curve being drawn above the line of zero voltage.

Position III shows the loop passing through the 180-degree region of its revolution in which position it is again parallel with the magnetic lines of force (flux), and for this instant no e.m.f. is being induced as is shown by the curve.

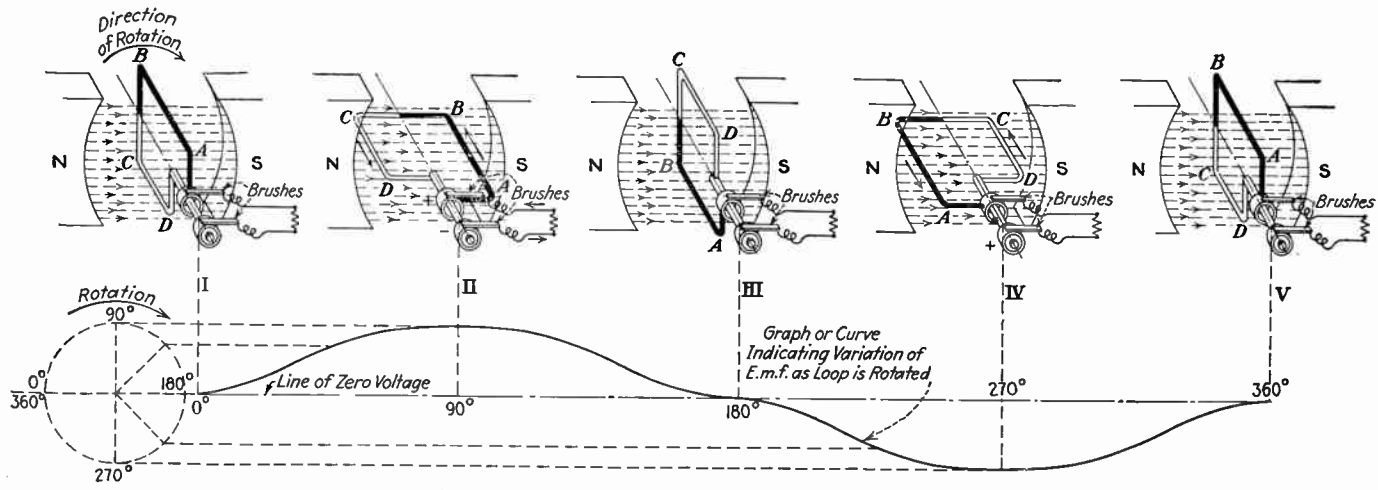


FIG. 49.—A simple illustration of the alternator principle.

Position IV is the 270-degree region of its revolution in which the conductor is once more cutting the field with the resultant current flow as shown by the curve. It is to be noted here that the sides *AB* and *CD* have completely reversed their position; that is, side *AB*, which was at the south pole in position II, is now at the north pole of the magnet. Therefore, the e.m.f. which has been generated in the loop during the first half of the revolution has now changed its direction, and is now indicated below the zero line of the curve.

Position V shows the loop as it is completing its revolution and is again parallel with the magnetic flux, the condition under which no e.m.f. is induced in the conductor. Sides *AB* and *CD* are now again in their starting positions, and, if the loop is again rotated in the same direction, another complete cycle of e.m.f. will be generated.

Any change in the direction of rotation, speed, or number of turns in the conductor, or in the strength of the magnetic field, will directly affect the value of the generated e.m.f.

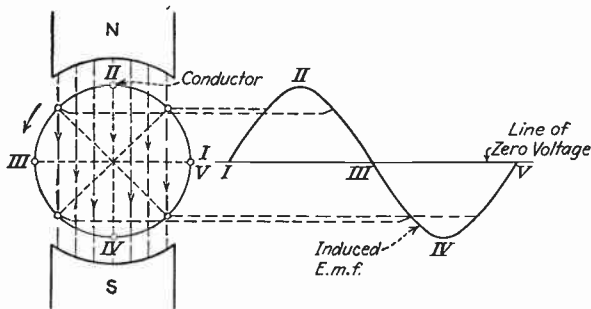


FIG. 50.—The angular relation between an armature conductor and the current it produces in an alternator.

Cycle and Frequency.—It is now evident that during the 360-degree revolution of the loop two alternations of current have taken place and flowed in the external circuit *R*. These two *alternations* of current constitute one *cycle* of alternating current, and the number of cycles per second is termed the *frequency* of the current.

Thus, an alternator is a machine for generating alternating current, or, in other words, a current which reverses its polarity periodically. The angular relation between the conductor and the cycle of current it produces is shown in Fig. 50.

42. Principles of the Transformer. *a. General Considerations.*—It has been shown that, if a moving magnetic field is developed in a coil by passing an alternating current through it, the moving field may be used to induce an e.m.f. into a near-by circuit, provided that the two circuits are magnetically related and in the proper axial positions. This condition is illustrated in Fig. 51(a) and is the fundamental theory upon which all transformers operate.

The transformer consists essentially of two separate coils of wire insulated from each other and wound upon a laminated iron core. The

winding which is connected to the alternating-current source is called the *primary* and that into which an alternating e.m.f. is induced is called the *secondary*. If it is desired to increase the voltage developed across the secondary winding relative to that impressed across the primary winding, the number of turns in the secondary must be more than those in the primary winding. This is called a *booster* or *step-up transformer*. If, on the other hand, it is desired to develop a lower voltage across the secondary winding than is put into the primary, then the number of turns in the secondary winding must be less than those in the primary. This is called a *step-down transformer*.

The transformation of voltage from low to high or vice versa is entirely dependent upon the turn ratio between the windings and not upon the size or character of the iron core. The core serves to complete the interlinkage of the magnetic lines of force between the two windings so as to insure the maximum transfer of energy (power) from one circuit to the other. The magnetic lines of force which do not interlink are called *leakage flux* or *leakage reactance*. The greater this leakage, the greater will be the losses and consequently the lower the efficiency. Other losses also contribute greatly to the inefficiency of transformers, namely, core losses and copper losses.

Core losses are due to the generation of heat in the core material produced by the molecular friction of the molecules when an alternating e.m.f. is applied to the core winding. These reversals introduce a power dissipation which represents an energy waste and takes away from the useful contribution of power from one circuit into another. These losses, known as *hysteresis*, may be reduced to a minimum by the use of special core material such as silicon steel or permalloy, in which the molecular friction is known to generate a minimum of heat. The losses may further be reduced by using the proper physical dimensions of core material to insure against hot operation.

Another serious core loss generally encountered in power transformers is that due to the generation of small circular whirls of magnetic lines of force known as *eddy currents*. These currents are set up in core materials because of the solidified cross-sectional structure of the core and are in a direction which opposes the magnetizing force of the primary winding. These so-called *eddys* may, however, be reduced to a negligible quantity by breaking up the cross-sectional structure of the core into sheets or laminations which are insulated from each other by special insulation varnish or colloid treatment. Hence, the magnetic lines of force which follow the plane of the core are in no way affected, but the small eddy current whirls are broken up into smaller components, thereby reducing the retarding effect that would otherwise be prevalent if a solid core structure were used.

Copper losses are reduced by the use of the proper size of wire to accommodate the input- and output-load conditions without over-heating.

b. Operating Conditions.—When the primary winding of a power transformer is connected to the alternating-current line, the value of the current flow in the primary winding will be affected by the value of the current flow in the secondary. If the secondary winding is operating under zero load, or on open-circuit conditions, a voltage will be developed by the secondary winding depending upon the ratio of turns, but no power will be developed in the winding. Hence the value of the current flowing

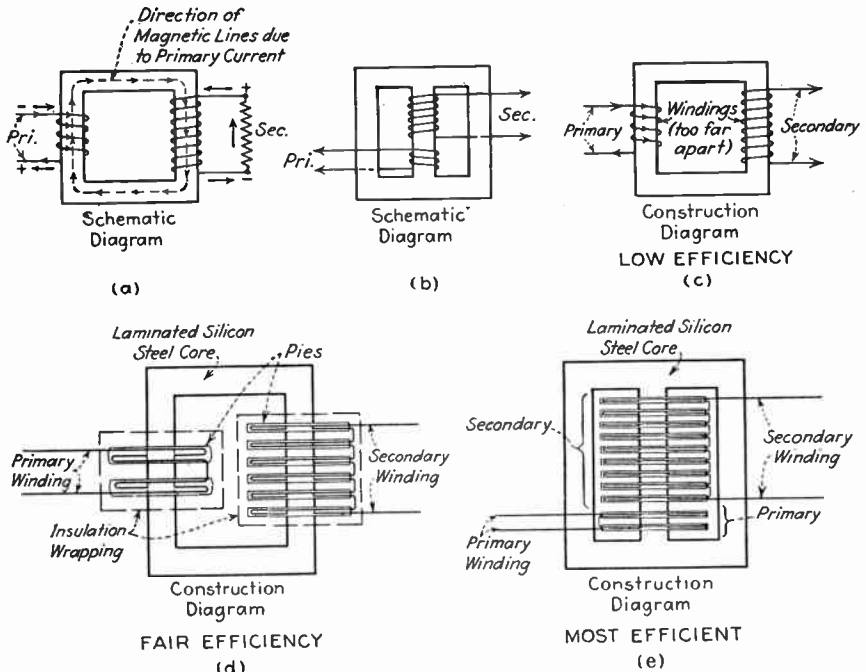


FIG. 51.—The electrical and constructional principles of the transformer.

in the primary circuit under no-load conditions is equal to $I = E_{eff}/Z$.¹ Actually, under these conditions, the current flowing in the primary winding will be very small owing to the fact that the voltages induced into both coils (self-induction) are in opposite directions. These induced voltages are, as a matter of fact, very nearly equal to the applied e.m.f.

Another way of looking at the circuit conditions when the secondary is unloaded is that very nearly all of the magnetic energy developed in the magnetic field around the primary winding is returned to the supply circuit, and consequently very little power is lost.

¹ The effective voltage E_{eff} is that which exists owing to the difference between the applied E_1 and the counter e.m.f., which is, obviously, very small.

Under loaded secondary conditions, however, the result is quite different. When the secondary circuit is closed, a current will flow in the winding which will develop a magnetic field about it. This field is 180 degrees out of phase with that produced by the primary winding, and consequently acts to oppose it, owing to the secondary voltage being 180 degrees out of phase with the applied primary voltage.

Now, since the current flowing in the secondary winding flows in such a direction as to oppose the primary magnetic lines of force that are producing it, the total lines of force (ϕ), produced by the primary, will be reduced, thereby increasing the current flow through the primary winding.

In other words, the opposing effect of the magnetic lines of force around the secondary against the primary effectively reduces the self-inductance and the counter e.m.f. of the primary, thereby resulting in an increase in the primary current. From this it would seem that a heavily overloaded secondary or short-circuited winding may actually reduce the primary reactance to a point at which an excessive and possibly damaging current may flow. However, this is not quite true in the actual transformer owing to the ever present magnetic-leakage reactance in the windings.

That is to say, there are lines of force present around the primary winding which do not interlink with the secondary. The same is true around the secondary. Now, since these lines in the respective windings do not react upon one another, consequently they cannot neutralize each other. This tends to maintain a certain amount of reactance in both the primary and secondary windings, and in effect is the same as if two additional reactance coils were connected in series with the primary and secondary circuits, respectively. Although this does tend to reduce the possibility of overloading the windings, it is imperative to maintain the reactance as low as possible to obtain a constant potential at variable loads, or, in other words, to improve regulation.

Now, since magnetic-leakage reactance represents lines of force which do not interlink with their associated windings, it is merely necessary to place the windings very close together to insure a maximum degree of mutual (M) inductance or coupling. This may be obtained by arranging the primary and secondary windings as illustrated in Fig. 51(b). Figure 51(c) illustrates the arrangement of the two windings which results in high-leakage reactance and poor voltage regulation.

In transformers where high step-up voltages are desired, the close proximity of the high-voltage winding to the exciting winding necessitates careful insulation to prevent voltage breakdown. This is generally accomplished by distributing the strain between turns through a sectional arrangement of the windings into layers or "pies," as illustrated in Fig. 51(d) and (e). Each pie is thoroughly insulated from the adjacent one by mica or micarta insulations, and the individual pies are wrapped

in specially treated empire cloth or insulating wrapping. All the pies are then connected in series to obtain the desired turn ratio between the windings.

c. Voltages, Current and Power Relations.—The voltage ratio developed between the primary and secondary circuits equals the turns ratio N between the two windings, namely,

$$N = \frac{E_p}{E_s} = \frac{N_p}{N_s} \quad (32)$$

The current I_p , flowing in the primary circuit, therefore, under load conditions will be

$$I_p = I_s \times \frac{N_s}{N_p} \quad (33)$$

and consequently, the secondary voltage, E_s , that is developed under load conditions will be

$$E_s = \frac{N_s}{N_p} \times E_p \quad (34)$$

Practical working formulas for any condition may thus be evolved from the preceding and following formulas:

$$E_p = \frac{E_s \times I_s}{I_p} \quad \text{volts} \quad (35)$$

$$I_p = \frac{E_s \times I_s}{E_p} \quad \text{amp.} \quad (36)$$

$$E_s = \frac{E_p \times I_p}{I_s} \quad \text{volts} \quad (37)$$

$$I_s = \frac{E_p \times I_p}{E_s} \quad \text{amperes} \quad (38)$$

Since the transformer is a device which transfers energy from one circuit into another, it is evident that the power which can be dissipated in the secondary winding with 100 per cent efficiency would be exactly the same as that flowing in the primary. Thus,

$$W_p = W_s \quad \text{watts}$$

However, since energy is lost in the heating effect of the wire I^2R , and core losses are present owing to eddy currents and hysteresis, the power output in the secondary circuit cannot be equal to that which is put into the primary. The efficiency is obtained by taking the ratio of the power output W_s to the power input, W_p , thus,

$$\text{Efficiency per cent} = \frac{W_s}{W_p} = \frac{\text{output}}{\text{output} + \text{losses}} \times 100 \quad (39)$$

Furthermore, since the power (W) expended in an alternating-current circuit is $W = E \times I \times p.f.$, then by applying this product to the transformer circuit we get, ($p.f.$ = power factor, see Sec. 62)

$$P_p = E_s \times I_s \times p.f.(s) \quad \text{watts} \tag{40}$$

$$P_t = E_p \times I_p \times p.f.(p) \quad \text{watts}$$

It may thus be seen from the various relations that the transformer may be designed for high-voltage or low-voltage outputs by merely obtaining the correct turns ratio between the two windings. The size of the wire has no bearing upon the voltage output. Hence, if it is desired to step up the voltage between the two windings 1,000 times, or say from 100 volts to 100,000 volts, the turns ratio must be 1:1,000. In other words, there must be 100 turns in the secondary winding to every 1 turn in the primary winding. Thus, if the secondary should require thousands of turns of wire to obtain this ratio, it would make no difference, in so far as ratio is concerned, whether a No. 18 B. and S. or an exceptionally fine No. 40 B. & S. gage wire were used. The voltage induced depends upon the turns ratio and not the size of wire used.

If, on the other hand, it is desired to develop power in the secondary circuit, then it will be necessary to have both the proper turns ratio and the proper size of wire to accommodate the amount of current flow in the circuit. If the wire is too small for a given amount of current flow, an I^2R drop results in the winding which causes the output voltage to be reduced. Furthermore, if the current is increased much beyond the rated carrying capacity of the wire, the winding will burn out owing to the overload condition. Also, it must be remembered, that, even though the transformer may not burn out, the development of *too much* heat in the windings lowers the efficiency of the transformer.

d. Relation of Frequency to Transformer.—The current I_p drawn by the primary of a transformer varies indirectly as the impedance Z_p of the primary. Furthermore, the impedance Z_p varies directly with the frequency f_p . Therefore, a given transformer designed, let us say, for 60-cycle operation has a primary impedance of 21 ohms when loaded. At 105 volts this transformer will draw 5 amp. under load. At 120 cycles the primary impedance is increased, let us say, to 42 ohms. The current drawn is then 2.5 amp., with no harm done to the primary winding.

Assume now that this transformer is connected to a 40-cycle line. The result is a lowered primary impedance, owing to the lower frequency, with a consequent rise in current. Let us say the primary impedance drops to 14 ohms. The current then rises to 7.5 amp., an increase of 50 per cent. Under these conditions the primary winding would overheat and might possibly burn out unless it was designed for the overload.

It is obvious then that a transformer will not be harmed if it is used at a higher frequency than that for which it is designed, whereas it will be damaged if used on a lower frequency unless an overload tolerance is allowed for in the primary winding.

The frequency of the alternating current delivered by the secondary of a transformer is always the same as that impressed across the primary.

The transformer as used in rectifier units is discussed further in the chapter on Rectifier Units (Chap. XIII).

II. INDUCTANCE

43. Induction.—When a current is first sent through a conductor, the lines of force around the conductor must be visualized as an expanding or moving magnetic field which expands until the apparent maximum extended position has been reached. This all happens in a fraction of a second. If the current is suddenly turned off by opening the circuit, the magnetic field will contract or collapse and totally disappear. This expanding and contracting effect of the magnetic field is the basis of all electromagnetic induction between fixed circuits, and takes place only when the magnetic field is moving (expanding and contracting). Obviously the magnetic field moves only when the applied e.m.f. or the resistance of the circuit is varied, as in the case of interrupted direct-current or alternating-current circuits.

If a conductor is placed in the vicinity of another conductor carrying an e.m.f. of varying intensity, and consequently having a varying field, an e.m.f. will be induced in it because of its being cut by this varying magnetic field. The current in this conductor is called an induced current and is due to the phenomenon called *electromagnetic induction*. All induction due to the flow of a varying electric current is electromagnetic in character.

Let us imagine two coils *A* and *B*. A current of varying intensity is flowing through coil *A*. Coil *B*, being near coil *A*, has a current induced in it by induction. As the field in coil *A* dies down, the lines of force around coil *B* also die down. In so doing, they cut and induce an e.m.f. in coil *A*, so that a third e.m.f. is produced by induction. First, there is the original e.m.f. in coil *A*; second, the induced e.m.f. in coil *B*; and, third, a reinduced e.m.f. in *A* due to *B*. This phenomenon is called *mutual induction*.

44. Self-induction.—*a.* The lines of force around a coil, through which a varying current is flowing, will expand and contract as the current varies. As they rise and fall, the magnetic field around each turn of the solenoid cuts itself and many adjacent turns, and consequently produces an e.m.f. greater than for one turn, the amount of which is dependent upon the number of turns, the amount of current flowing through the turns, and the number of lines of force threading the coil. This action results in an induced e.m.f., which is in the direction opposite to the impressed e.m.f. of the coil, which tends to retard the flow. This opposing effect is present only when the field is expanding. This phenomenon

is called *self-induction*, because it is an e.m.f. induced in a conductor by its own moving magnetic field.

b. The *self-induction* of an electrical circuit is defined as *that phenomenon whereby an e.m.f. is induced in the circuit itself when the current in the circuit changes, or varies.* This important factor the student should understand thoroughly in order to comprehend clearly the function of inductances when they are connected in alternating-current circuits.

c. Referring again to Sec. 34 on Fields around Conductors; it was explained how a current passing through a wire creates a magnetic field about it, and how this field can be increased by coiling the wire into helical form and inserting into it an iron core. Let us consider what happens in the circuit itself when the magnetic field is either expanding or contracting through a variation of the current flow; remembering that, whenever a current varies, the magnetic field around the wire correspondingly varies (moves). It is extremely important to grasp the following facts: *When a magnetic field appears and disappears, it produces an e.m.f. in all conductors cut by it.* It is immaterial whether these conductors belong to a separate circuit, or whether they constitute the circuit (the helical coil referred to above) in which the magnetizing current is flowing.

d. The e.m.f. which causes the current to flow through the coil is called the *applied* e.m.f. There is also an e.m.f. which is produced by the circuit itself, independent of the applied e.m.f.; this is the *counter* e.m.f., and it is due to self-induction. For an example see Fig. 52. Here is a simple electric circuit, consisting of a battery, a switch, a resistance, and a coil. In this circuit a current is flowing continuously, whether the switch is open or closed; the difference being only in the strength of current, that is, when the switch is open, a minimum of current flows, owing to the resistance, and when the switch is closed, a maximum of current flows because the resistance is then short-circuited.

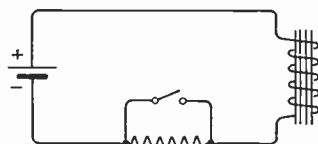


FIG. 52.—A simple electrical circuit to illustrate the effects of induction.

e. The student will readily see that in each of these situations there will be a magnetic field present about the coil. Hence, when the switch is open, the magnetic field of a certain density extends out from the coil. When the switch is closed, therefore, the field about the coil will increase in density. In other words, the field has actually expanded from a minimum to a maximum density when the switch is closed and from a maximum back again to a minimum when the switch is opened.

Now, it is quite apparent that if the switch should constantly be opened and closed, there would be a moving magnetic field, expanding and contracting at a definite rate, depending upon the speed with which the circuit is opened and closed. What effect would this action have on the circuit itself? This is one of the most important principles associated with all forms of electric circuits in which the current is changing or varying.

When a current is starting to flow in a conductor, due to an applied or impressed e.m.f., the magnetic field about the conductor is expanding and becoming denser. During this period of expansion, the field, consisting of magnetic lines of force, is cut through by the wire or coil by which it is produced. This causes a second current to be set up in this conductor, but in an opposite direction to that of the applied or impressed e.m.f. Therefore, during the period of field expansion, the coil or wire acts as an opposition to the applied e.m.f. and thus prevents the current in the circuit from reaching its maximum until the cutting of the field by the conductor has been

completed. This continues usually for a fraction of a second only, but the significance of this result is readily seen when the circuit is made and broken at a great rate of speed.

Thus far, only the results of a starting or expanding field have been noted. When the circuit is broken and the current is decreasing, the process is reversed; the field as it contracts is again cut by the coil, but, as this cutting is now in the opposite direction, another self-induced e.m.f. is produced, which is now in the same direction as the impressed e.m.f. and which tends, therefore, to prolong the duration of current. The value of the back pressure at make (when the circuit is closed) is never greater than the impressed e.m.f., otherwise the current could not rise in value. Its value at break (when the circuit is opened), however, is not limited by any such condition, and, therefore, when the break is very sudden, its value at that instant may be very great.

f. Again, the magnetic field around a steady current represents a certain amount of energy stored up in the surrounding space. For example, it may be said that the energy was supplied at the expense of the current in the circuit when the e.m.f. was started, and that this same energy was returned to the circuit when the current was stopped. The starting of a current in a circuit, therefore, resembles, in a logical manner, the starting of some heavy mass, a grinding wheel, for example. It cannot be started suddenly, nor does it naturally stop suddenly. Here the starting of the wheel would be analogous to a force overcoming the inertia of the mass; in electrical terms the highly inductive circuit is analogous to a body of great mass. Referring again to the wheel, the amount of energy associated with it when in steady motion, and the magnitude of all the effects depending on that energy, will vary with its mass. This example illustrates the effect produced by induction in a circuit, and any conductor or coil producing these effects by virtue of its inductive properties is called an inductance. *Self-induction* in a circuit, therefore, tends to prevent a change in the strength of the current flowing through it. This is sometimes called the *flywheel effect*.

g. Hence it can readily be seen that the greater the number of turns in a coil, or, even more so, the greater the amount of iron in the core of a coil, the greater will be its *self-inductance*, and therefore, the greater the self-inductance of any circuit the greater will be its ability to oppose the e.m.f. impressed upon it. From this it can clearly be understood that there might be a possibility of a circuit having a self-inductance so large that no current of a varying character could pass through it.

From the electron viewpoint the self-inductive effect in a circuit due to a changing or varying e.m.f. may be pictured as the accelerating action of each electron in the conductor urging the neighboring electrons in the opposite direction. The electrons, therefore, as small bodies or masses follow the general law of mechanics in which the force tending to overcome a mass (set it in motion) will be opposed by the mass just as the grindstone opposes, due to its mass, the applied force during the period in which it is being set into motion. All of the electrons that are being accelerated in the conductor, therefore, resemble a large quantity of "little masses," and consequently as more of them are progressively accelerated their tendency to oppose the flow along the entire route of the conductor will be greater. This explains the reason for the ability of coils with iron cores to have a great effect upon the flow of current during the period in which the magnetic field, around the conductor and the iron core, is being expanded. That is to say, as each electron in the magnetic material is being accelerated by the electron in the conductor, its force (the magnetic field force) will tend to oppose the progressive movement of the electron in the conductor.

h. Lenz's Law.—The summarizing of Lenz's law as applied to self-induction may, therefore, be defined as that phenomenon in which an e.m.f. is induced in the circuit itself during the period in which the elec-

trons in the circuit are accelerated or coming to a stop. This e.m.f. is called the *counter e.m.f.* due to self-induction, and it always produces a magnetic field which opposes any change in the force which produces it. Counter e.m.f. is always present in a circuit when the current in the circuit is changing or varying and is proportional to the rate at which the current is varying.

45. Mutual Induction.—(a). Figure 53(a) illustrates two coils *A* and *B* placed in an axial relationship to one another. Coil *A* is arranged with a switch and resistance *R* for varying the rate of current flow without actually opening the circuit. Assuming a steady flow of current through coil *A* with the short-circuiting switch open, coil *B* is brought into close relationship with the steady magnetic-field condition surround-

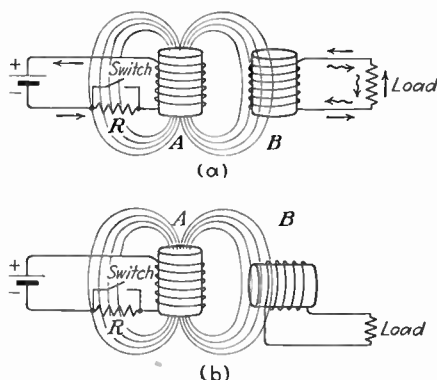


FIG. 53.—Two coils in axial relation to show effect of mutual induction.

ing it, because of the non-varying amplitude of the current flow. The free electrons in coil *B*, although in a state of constant agitation due to temperature, are not moving progressively from point to point and consequently do not constitute a flow of electric current.

b. If now the switch is closed, short-circuiting the resistance *R*, the flow of the electrons through the coil *A* will be increased, causing a sudden accelerating action of the electrons in the circuit. This will result in a momentary expansion of the magnetic field about coil *A* and a pulse movement in the space surrounding the coil. Now, since coil *B* is assumed to be located within this area, the free electrons within this coil will be subjected to the moving magnetic forces about the coils and will be displaced in a definite direction as indicated by the wavy arrows. In other words, the current flow in the coil *B* (the induced current) will be in an opposite direction to that in the coil *A* during the period in which the current in the coil *A* is increasing. The straight arrows indicate the direction of the induced current during the period in which the current in coil *A* is decreasing.

c. As soon as the field has expanded and ceases to move, the electrons in the circuit of coil *B* will cease to move progressively and the current flow will stop. If now the switch is opened, the number of free electrons moving progressively in circuit *A* will be decreased, resulting in the contraction of the magnetic field. The electrons in

circuit *B*, however, are again subjected to a movement of the magnetic field and are urged to move in an opposite direction to that in which they were displaced when the field was expanding. Hence, if the switch in circuit *A* is constantly open and closed, an alternating e.m.f., and an alternating current, will be induced in circuit *B*.

d. The direction of the induced current in circuit *B* will be opposite to that of the current flow in circuit *A*, due to the mass effect of the electrons in the circuit during the period in which they are being set in motion. Hence, the current flow in the circuit *B* will be in the same direction as that in circuit *A*, only during the period in which the current flow in circuit *A* is decreasing. *Current flow is present in circuit B only during the period in which it is cutting or being cut by moving magnetic lines of force.* The term *cut* or *cutting* refers to the action of the magnetic lines of force upon the free electrons in the conductor and depends upon the *movement* of the field or coil and the position of the coil with respect to the lines of force.

e. Figure 53(b) illustrates a position of two coils *A* and *B* in which no e.m.f., and consequently no current, is induced into circuit *B*, even though the magnetic field about the coil *A* is assumed to be moving. This is due to the fact that the lines of force are parallel with the turns in the conductor, and therefore they do not displace any electrons in the circuit *B*. For further information on this action see explanation of Fig. 303.

f. Let us assume the coil *B* to be in the proper magnetic relation to coil *A*, and the field about coil *A* to be expanding and contracting due to a varying e.m.f., such as is produced by opening and closing the switch in rapid succession. An alternating current will flow in the circuit *B* because of the constant displacement of electrons in that circuit. This will result in the creation of a magnetic field around coil *B* because of the continuous movement of electrons in the circuit.

g. However, since these electrons are first displaced in one direction and then in the other, the polarity of the magnetic field around the coil *B* will be constantly changing, the number of changes depending upon the rate of interruption in the inducing circuit *A*.

h. The magnetic field now present about the coil *B* will have an effect upon the moving electrons in the circuit *A* and, consequently, its magnetic field. This results in a constant action and reaction of both magnetic fields upon one another during the period in which the current in any one circuit is changing or varying. It is this effect of the magnetic reactions between circuits which is called *mutual induction*, and all circuits which are so related are referred to as magnetically coupled circuits.

i. The effect of induction is a measurable property of a circuit and is called *inductance*. The physical unit (coil or conductor) in the circuit causing this effect is called or referred to as an *inductance*. The term inductance may, therefore, mean an abstract property of a circuit or a component part of the circuit.

j. The effects of magnetic coupling reactions between circuits are treated more thoroughly under Transformers (Sec. 42) and Oscillatory Circuits (Sec. 151).

46. Inductance Measurements.—The ability of a coil to function as a storehouse of electrical energy in magnetic form is expressed in electromagnetic units measured in centimeters which for practical purposes are called *henrys*, in honor of the discoverer of electromagnetic induction, Joseph Henry (1797–1878). Inductance or self-inductance is usually denoted by the letter *L* and is an expression of electrical mass or ability to store up energy in electromagnetic form.

This is dependent upon the number of turns in a coil, the diameter of the coil, the length of the coil, the character of the core, and the rate of

current change in the coil. Hence the inductance (self-inductance) of a coil can be calculated by the general expression

$$L = \frac{N\phi}{I} \quad (41)$$

where N = the number of turns in the coil.

ϕ (phi) = the total number of lines of force linkages with one turn.

I = the varying current flow in the coil.

A coil is said to have a value of 1 henry of inductance if a varying current of 1 amp. per second induces an e.m.f. of 1 volt in the circuit. Similarly, if two circuits are magnetically coupled, the mutual inductance will be 1 henry if a varying current of 1 amp. per second induces an e.m.f. of 1 volt into the second circuit. Thus, since the self-inductance of a coil is proportional to the square of the number of turns, the mutual inductance of two inductances will, therefore, vary as the square root of the two inductances, or

$$M = \sqrt{L_1 \times L_2} \times k \quad (42)$$

[For value of k see Eq. (51).] Also the induced voltage may be computed by the ratio

$$E = M \frac{I_A}{t} \quad (43)$$

However, since the electrons in the secondary circuit accelerate in an opposite direction to those in the primary circuit, the induced e.m.f. and current will be out of phase. Hence to indicate this condition in the formula a small - (minus) sign should be inserted, thus

$$E = -M \frac{I_A}{t}$$

In other words, this serves to indicate that the e.m.f. in the secondary circuit is in a direction which tends to oppose that which is producing it. Similarly, if the counter e.m.f. due to self-induction in a single circuit is to be determined, simply change the symbol from M to L , thus

$$E = -L \frac{I_A}{t} \quad (44)$$

These formulas are theoretical and do not apply to practical designing.

47. Practical Inductance Formulas.—The self or mutual inductance of a wire or coils may be calculated from their respective geometrical dimensions by many formulas. Some of the more important ones directly applicable to practical design will follow. All calculations are expressed

in centimeters, since it will be a simple matter to convert these values into inches and henrys as follows:

1 centimeter = 0.3937 inch or 0.01 meter (1 in. = 2.54 cm.).

1 henry (h) = 1,000,000,000 centimeters, or 10^9 cm.

1 millihenry (mh) = 1,000,000 centimeters, or 10^6 cm.

1 microhenry (μ h) = 1,000 centimeters, or 10^3 cm.

1 centimeter of inductance = 0.000000001 henry.

1 henry = 1000 mh. = 1,000,000 μ h.

To convert centimeters into microhenrys divide by 1,000 or multiply by 10^{-3} .

1. *For Parallel Wires.*—For a circuit in which the current flows in two parallel wires but in opposite directions, the magnetic fields react in a manner in which both fields may cancel or neutralize.

$$L_{eff} = L_1 + L_2 \pm 2M \quad (45)$$

where L_1 and L_2 = the self-inductance in centimeters of the two wires, and M = the mutual inductance in centimeters between them.

2a. *For a single-layer Coil (Air Core).*

$$L = 4\pi^2 r^2 n^2 l K \text{ cm. (Nagaoka's Formula)} \quad (46)$$

where r and l (radius and length) are expressed in *centimeters*.

n = number of turns per centimeter.

K = constant factor determined by the ratio of d/l (see Table IIA).

If the total number of turns of wire are used, instead of the number of turns per centimeter, the formula may be expressed

$$L = \frac{4\pi^2 r^2 n^2 K}{l} \text{ cm.} \quad (47)$$

$$(\text{cm.} \times .001 = \mu\text{h})$$

where N = the total number of turns.

This formula is based upon the assumption that the spacing between the turns is infinitely small, that is to say, the turns may actually touch without making electrical contact. If, however, insulation or space-wound coils are required, then a correction factor must be used in order to obtain accurate results.

To make the necessary correction for this condition, subtract the quantity Δ from the result of the previous formula.

$$\Delta = 2\pi dn l [A + B] \quad (48)$$

where n = number of turns per centimeter.

A = ratio of d_0/D_1 .

d_0 = diameter of the bare wire.

D_1 = diameter over the insulation or the distance between the centers of consecutive turns.

B = the total number of turns on the coil.

b. For rapid and fairly accurate calculation of the inductance of a single-layer air-core coil, the following formula may be used:

$$L = \frac{0.067 \times d^2 \times N^2}{d + 3l} \text{ microhenrys} \quad (49)$$

where d = the diameter of the coil in inches.

l = the length of the winding in inches.

TABLE IIA.—VALUE OF K FOR NAGAOKA'S FORMULA ($L = 4\pi^2 r^2 n^2 K$ cm.)
(For helical-wound coils)

Ratio of d/l	K	Ratio of d/l	K	Ratio of d/l	K
0.05	0.9791	1.9	0.5379	5.6	0.2981
0.1	0.9588	2.0	0.5255	5.8	0.2916
0.2	0.9201	2.2	0.5025	6.0	0.2854
0.3	0.8838	2.4	0.4816	6.2	0.2795
0.4	0.8499	2.6	0.4626	6.4	0.2739
0.5	0.8181	2.8	0.4452	6.6	0.2685
0.6	0.7885	3.0	0.4292	6.8	0.2633
0.7	0.7609	3.2	0.4145	7.0	0.2584
0.8	0.7351	3.4	0.4008	7.2	0.2537
0.9	0.7110	3.6	0.3882	7.4	0.2491
1.0	0.6884	3.8	0.3764	7.6	0.2448
1.1	0.6673	4.0	0.3654	7.8	0.2406
1.2	0.6475	4.2	0.3551	8.0	0.2366
1.3	0.6290	4.4	0.3455	8.5	0.2272
1.4	0.6115	4.6	0.3364	9.0	0.2185
1.5	0.5950	4.8	0.3279	10.0	0.2033
1.6	0.5795	5.0	0.3198	11.0	0.1903
1.7	0.5649	5.2	0.3122	12.0	0.1790
1.8	0.5511	5.4	0.3050	13.0	0.1692

Example 1: Find the inductance of a single-layer coil of 10 turns wound on a cylinder of 1.5 in. in diameter, the length of the coil being 0.61 in. Proceed as follows:

$$L = \frac{0.067 \times 1.5^2 \times 10^2}{1.5 + 3 \times 0.61} = 4.52 \mu\text{h}$$

Comparing this result with the extremely accurate Nagaoka formula, we find

$$L = 4.4 \mu\text{h. (Nagaoka)}$$

compared with

$$L = 4.52 \mu\text{h.}$$

Example 2: To find the inductance of a single-layer coil having a diameter of 4 in., 100 turns, and a length of 2.2 in.,

$$L = \frac{0.067 \times 4^2 \times 100^2}{4 + 3 \times 2.2} = 1,010 \mu\text{h.}$$

The same coil calculated from the Nagaoka formula (46) was found to be 1,000 μh .

A set of curves based on Nagaoka's inductance formula is reproduced here as Fig. 54, by permission of the publishers, from the April, 1931, issue of *Electronics*.

3. *For Iron-core Coils (Closed Core).*—The previous calculations have been entirely for single-layer coils without magnetic cores, in which case the inductance depended principally upon the geometrical dimensions, with a small correction factor depending upon the frequency or rate of current change through the coil.

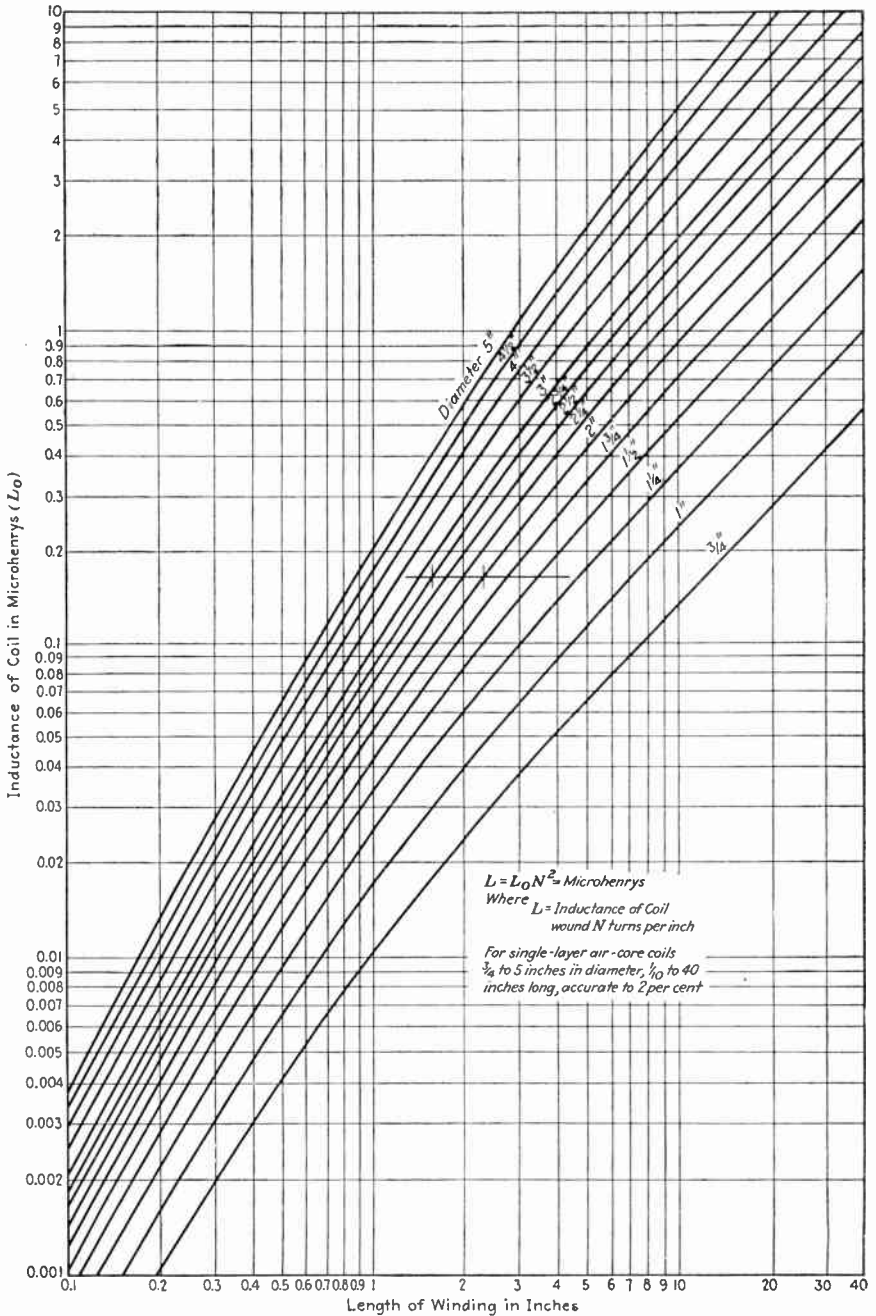


FIG. 54.—Curves based on Nagaoka's inductance formula. (Courtesy of Electronics.)

For iron-core coils, however, the inductance L depends on the physical dimensions of the coil and the physical properties of the iron, and, since these properties are variable, the calculations will be more difficult.

The inductance of an iron-core coil, therefore, depends upon the permeability of iron, and this will vary over the cross section of the core since the magnetizing force is not constant over that area. Also, if the current in the coil producing the magnetizing force varies, the value of the magnetic permeability μ will change. Consequently, some average value of the permeability must be obtained in order to make definite computations.

Perhaps one of the most simple formulas that can be derived for iron-core coils, if we assume the most general conditions, namely, that of a closed iron core with a small air gap, is

$$L = \frac{0.4N^2\mu A}{l} \quad (50)$$

where L = inductance in henrys.

l = length of the air gap in centimeters.

A = area of the surface of the iron core at the air gap.

$\mu = 1$.

48. Mutual Inductance Measurements.—*a.* When coils are connected in series so that their respective magnetic fields do not react upon one another, the effective or total inductance would be the algebraic sum of the separate inductances, expressed as follows:

$$L = L_1 + L_2$$

b. If the coils are placed close enough to cause magnetic reaction between them, then

$$L = L_1 + L_2 - M_{1,2} = \pm M_{2,1}$$

where $M_{1,2}$ is the mutual inductance or the e.m.f. induced across coil 1 by the rate of the current change in the second coil 2 (1 amp. per second) and $M_{2,1}$ is the e.m.f. induced into coil 2 by the current change in coil 1. Hence, the e.m.f. reactions between the coils 1 and 2 result in

$$L = L_1 + L_2 \pm 2M$$

Now, if the coils are so connected that the magnetic fields assist each other, then

$$L_a = L_1 + L_2 + 2M$$

Consequently, if the windings of one coil are changed, so that the current flows in a different direction in one coil than in the other coil, the magnetic fields oppose and react as follows:

$$L_o = L_1 + L_2 - 2M$$

Hence,

$$L_a - L_o = 4M$$

and thus

$$M = \frac{L_a - L_o}{4}$$

c. The degree of magnetic coupling reactions, or the coupling coefficient k , as it is generally called, may thus readily be computed by the ratio

$$k = \frac{M}{\sqrt{L_1 \times L_2}} \quad (51)$$

where M represents the mutual inductance in henrys.

$$M = \sqrt{L_1 \times L_2} \times k \quad (42)$$

The value of the coupling coefficient k will lie somewhere between 0.5 and zero if the coils are widely separated (loosely coupled), and above 0.5 if closely related (tightly coupled).

III. CAPACITY

49. The Electrostatic Field and the Condenser.—When two metal plates are separated by air or some other insulating material, an electric field will be manifest when the two plates are connected to a source of e.m.f. This is apparently due to a displacement of the electrons in the insulating material (dielectric) in the direction of the positive plate and a similar displacement of the nucleus toward the negative plate. This action, however, is not due to free electrons, since all insulators are believed to have very few or no free electrons in their structure. Consequently the separation of charges in insulating materials may be visualized as an area of polarized atoms in which the electrons in the outer orbits of each atom strain toward the positive plate and the inner electrons and nucleus strain in the direction of the negative plate. In other words, the orbital electrons are assumed to be separated from their nucleus but are still under the attractive influence of the nucleus.

This hypothesis may be substantiated from the evidence that a meter connected in series with the two plates will show a progressive movement of electrons (current flow) through the entire circuit at the instant the circuit is closed which immediately thereafter decreases gradually back to zero. Now, although the circuit is still closed, there is no longer any indication of current flow, and yet, if the applied e.m.f. to the plates is disconnected and a pressure-indicating meter (voltmeter) connected across the plates, a discharge deflection will be noted. This definitely indicates that the area between the plates had been strained and established a difference of potential, which, when connected to the voltmeter, was relieved in moving the circuit electrons through the voltmeter coil and its resistance.

Obviously, if the initial closing of the circuit indicated only a momentary or displacement current, no free electrons could have been present in the dielectric material, otherwise the current flow would have been continuous as in the case of simple conducting circuits. Hence, the electrons in the external circuit of Fig. 55(a) will flow only during the period in which the electrons in the dielectric are displaced, and once they have obtained this position the circuit current will cease to flow. If, however, the polarity of the applied e.m.f. is reversed, the electrons in the dielectric will be displaced toward the opposite plate since the positive charge is now on that plate and the electrons in the entire circuit will again be displaced, but now in the opposite direction. Consequently,

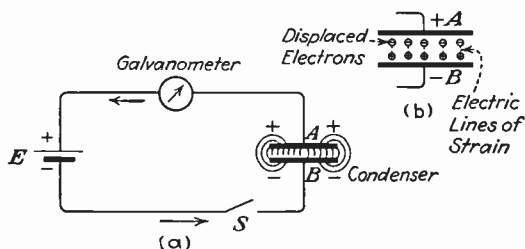


FIG. 55.—Simple circuit to show charging of a condenser.

if the applied e.m.f. across the two condenser plates is of an alternating character such as is produced by an alternating-current generator or transformer, the electrons will be continuously displaced, first in one direction and then in the other. Actually the circuit electrons will not move around the circuit progressively but will merely shift from left to right, as the electrons in the dielectric are displacing first in one and then in another direction.

It can readily be seen that the charging of a condenser is the process of raising the potential of one plate to a higher point with respect to the other; or, in other words, when the electrons in the dielectric are displaced toward the positive plate so that there is a separation of charges in the dielectric, the condenser is said to be charged.

This condition is generally represented by vertical lines between the plates as indicated in Fig. 55(a) to show that a state of strain exists between the plates when an e.m.f. is applied. If the charged condenser is disconnected from the charging e.m.f. and shunted by a high-resistance conductor, the potential difference due to the polarized atoms in the dielectric material (displaced electrons) will cause the electrons in the shunt circuit to be moved in the direction of the positively charged plate, thus reducing the potential difference by restoring the polarized atoms to their normal, or uncharged, position. When this takes place the condenser is said to be discharged.

Summarizing, it is seen that the charging process of a condenser as illustrated by Fig. 55 is one in which, when the switch is closed, electrons will be displaced in the dielectric in the direction of plate *A* due to the application of a steady potential from the source *S*. This results in a movement of electrons out of the plate *A* into the circuit, thereby displacing electrons throughout the entire circuit. These displaced electrons try to enter at the plate *B* and accumulate until the potential difference between the plates is equal to the e.m.f. of the source. If an excessive accumulation should take place at this point, the condenser would discharge until the charge is equal to the source e.m.f. Hence, the charges in a dielectric will cease to move when the potential difference between the plates equals the applied e.m.f., and, if the applied potential is constantly maintained, no electrons (current) will flow in the circuit.

50. Capacity of a Condenser.—The number of electrons necessary to establish a difference of potential between two plates is called the *capacity*. That is to say, the more electrons displaced in the dielectric the greater the capacity of a condenser.

The unit of electrostatic capacity is called the *farad*. A condenser which has a capacity of 1 farad would require 1 coulomb of electricity to bring its plates up to a potential difference of 1 volt. However, since the physical dimensions of a condenser with that capacity would involve a tremendous plate area, it is not used for practical purposes. In electrical and radio engineering practice the values of capacity are generally always expressed in micro-units, or microfarads. The microfarad is 1/1,000,000 part of 1 farad, consequently a condenser having a capacity of 1 mf. would require a charge of 1/1,000,000 of 1 coulomb to charge it up to a potential difference of 1 volt. Stated in another way, a current of 1 microampere flowing for 1 sec. would charge the plates to a potential difference of 1 volt.

The capacity of any condenser is measured by the quantity of electricity with which it must be charged in order to raise its potential from zero to unity. In other words, a small capacity would require only a small quantity of electricity to raise its potential from zero to unity while, on the other hand, a large capacity requires a greater quantity of electricity to establish a potential increase from zero to unity. Thus it can be clearly seen that the potential of a condenser depends upon the quantity of charge and its capacity:

$$C = \frac{Q}{E} \quad (52)$$

or

$$Q = CE \quad (53)$$

The capacity of a condenser is increased by one or more of the following methods:

1. Increasing the area of the metal plates.
2. Decreasing the distance between the plates (thinner dielectrics).
3. Increasing the dielectric constant of the insulating material.

51. Specific Inductive Capacity.—The specific inductive capacity or the dielectric constant of various insulating materials is the ability of such materials to exhibit electric charges on the sides facing the metal plates of a condenser.

Let us assume a condenser made up of two plates separated by an air spacing of $\frac{1}{32}$ in., and let the quantity of electricity required to charge it to 1 volt be measured. Then let $\frac{1}{32}$ -in. mica or glass plate be inserted in place of the air dielectric and the quantity of electricity again measured. It will now be found that the quantity of electricity is somewhere between five and ten times as much as when the air dielectric separated the plates.

It is this ratio of the capacity of a condenser when other than air is used, to that if air or a vacuum is used, which is called the *specific inductive capacity* of a dielectric.

The following table gives some values of specific inductive capacities of various insulating dielectrics relative to air.

TABLE III.—SPECIFIC INDUCTIVE CAPACITIES

Substance	Dielectric Constant <i>K</i>
A vacuum	0.94
Hydrogen	0.9997
Air at 0°C.	1.000586
at 20°C.	1.000576
Carbon dioxide	1.0008
Gases	1.00007–1.009
Liquid oxygen	1.478
Oil	2.2–4.7
Paraffin wax	1.99–2.29
Sulphur	2.24–4.2
Ebonite	2.6–3.48
Rubber	2.0–3.5
Paper, dry	2.0–2.5
Fiber	2.5–5.
Shellac	2.74–3.73
Gutta-percha	3.3–4.9
Quartz	4.49–4.55
Bakelite	4.5–5.5
Glass, various grades	5.0–10.0
Mica	2.5–6.7
Porcelain	5.7–6.8
Slate	6.6–7.4
Water	81.0

Water has an exceptionally high dielectric constant but cannot be used because of its electrical conductivity due to ionization.

52. Energy in a Charged Condenser.—The amount of energy or work done in moving the electrons in the dielectric medium between the plates of the condenser so as to establish a difference of potential is expressed

in joules. This represents an expenditure of energy which may be readily calculated from the formula

$$W = \frac{1}{2}CE^2 \text{ joules} \quad (54)$$

Thus, when a condenser is charged, the potential energy is supplied by the source e.m.f., and, when it is discharged, its potential energy is converted into kinetic energy (the motional energy representing the electrons in motion).

53. Capacity of Condensers in Parallel.—Figure 56 represents a common arrangement of condensers in parallel.

Assuming each condenser to have an individual capacity of 0.002 mf., the total effective capacity will be 0.006 mf. This can readily be seen to be so owing to the larger combined electrode area of the condensers. Let the letter Q represent the total charge given to them; then a difference of potential of E units between the points X and Y , which are the terminals of the systems, is produced.

The charges in the condensers will be greater, therefore, owing to the increase of plate area, but the E will be the same for all of them.

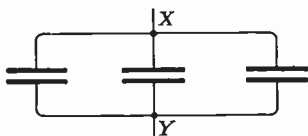


FIG. 56.—Three condensers in parallel.

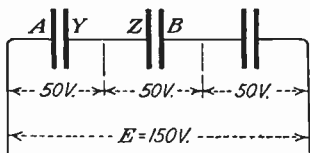


FIG. 57.—Three condensers in series.

Thus, if the charges q of each condenser are added,

$$Q = q_1 + q_2 + q_3 \quad (55)$$

Hence, if the total capacity of three condensers connected in parallel is the result of the addition of the three charges, the formula for condensers in parallel will obviously be

$$C = C_1 + C_2 + C_3 \quad (56)$$

54. Capacity of Condensers in Series.—Figure 57 represents three condensers connected in series. Assuming each condenser to have a capacity of 0.002 mf., the total effective capacity under this arrangement will be less than the lowest of the component capacities.

The formula for condensers of equal capacities connected in series is expressed as follows:

$$C = \frac{c}{n} \quad (57)$$

where n represents the number of condensers and c the capacity of one condenser.

Hence, in the above example,

$$C = 0.0006\frac{2}{3} \text{ mf.}$$

If the condensers are of unequal capacities, however, the formula reads as follows:

$$C = \frac{1}{\frac{1}{c_1} + \frac{1}{c_2} + \frac{1}{c_3}} \quad (58)$$

Example: Three condensers having capacities of 0.002, 0.004, 0.002 mf. are connected in series. What is the total effective capacity?

Solution:

$$C = \frac{1}{\frac{1}{c_1} + \frac{1}{c_2} + \frac{1}{c_3}}$$

or

$$C = \frac{1}{\frac{1}{0.002} + \frac{1}{0.004} + \frac{1}{0.002}} = 0.0008$$

Answer: Total effective capacity = 0.0008 mf.

It will be seen that whenever unequal capacities are connected in series, the total effective capacity will always be less than the smallest condenser value in the circuit.

Why the capacity is reduced when condensers are connected in series will be understood by analyzing the dielectric effect in the two condensers of Fig. 57.

The plates *Y* and *Z*, being electrically connected, actually form a single plate. Assuming the dielectrics of each capacity to have a like thickness, it can readily be seen that by joining both dielectrics in a series formation, the distance between plates *A* and *B* is merely widened and, consequently, the thickness of the dielectric is increased. If each of two condensers has a capacity of 0.002 mf., therefore, and is connected in series, the dielectric thickness is doubled and results in a reduction of capacity to one-half of its former value, or 0.001 mf.

From this it can be seen that the capacity of *n* number of condensers connected in parallel and charged from the same source would represent a total energy of

$$W = \frac{1}{2}E^2 \times n \times C \quad \text{joules} \quad (59)$$

Evidently, each constituent condenser is charged to the same extent as a single condenser, and the combination has, therefore, *n* times more energy, or, in other words, the parallel combination is equivalent to one condenser of *n* times greater capacity.

If a number of condensers, *n*, are connected in series, the capacity is *C/n*, and, consequently, the total energy is

$$W = \frac{1}{2}E^2 \frac{C}{n} \quad \text{joules} \quad (60)$$

Now, as the total energy is equally shared by the constituent condensers, each must possess $1/n^2$ of the amount that each single condenser must

have. Also, if we consider that the potential difference between the plates of each condenser is E/n , then the individual energy stored between the plates of each condenser is

$$W = \frac{1}{2} \left(\frac{E}{n} \right)^2 \times C \quad \text{joules} \quad (61)$$

55. Capacity Measurements.—It has been shown that the capacity of a conductor or condenser is defined and measured by the quantity of electricity required to raise its potential from zero to unity. The capacity given by this expression is the capacity with reference to the body or bodies at zero potential from which the potential of the given conductor or condenser is raised by the charge.

The general formula for determining the approximate capacity of a condenser consisting of two parallel plates separated by any form of dielectric may be expressed as follows:

$$C = \frac{KA}{4\pi d} \quad \text{electrostatic units or centimeters} \quad (62)$$

where A = area of one metal plate in square centimeters.

d = distance between the plates (thickness of the dielectric) in centimeters.

K = the dielectric constant (see Table IIB, preceding).

C = the capacity in centimeters.

9×10^5 cm. = 1 mf.

A more accurate expression of the preceding formula may be obtained by considering the number of dielectric sheets between the metal plates:

$$C = \frac{nKA}{4\pi d} \quad \text{centimeters} \quad (63)$$

where n = the number of dielectric sheets between the plates.

These formulas cannot give a very accurate result regardless of the care exercised in making the physical measurements because of the curvature of the electrostatic field between the outside edges of the plates. For practical purposes, however, the calculations become sufficiently accurate for approximate determinations of capacity.

For the calculations of approximate capacity in variable plate condensers the formula is relatively more complicated:

$$C = Kn \frac{\pi(r_1^2 - r_0^2)}{4\pi d} \times \frac{\theta}{360} \quad (64)$$

where r_1 = the radius of the smaller sets of plates, r_0 the radius of the fixed plates, n the number of dielectric sheets between the plates, and θ the angle in degrees through which the rotary plates have been turned from the position of minimum capacity.

The formula may be greatly simplified by assuming the use of an air dielectric variable condenser with the entire number of rotary plates meshed in a position of maximum capacity. Then the total capacity may be calculated from the formula

$$C = \frac{2nA}{4\pi d} \quad \text{centimeters} \quad (65)$$

where n = the total number of rotary plates.

d = the thickness of the dielectric between one stator and one rotary plate in centimeters.

A = the area of one rotary plate in centimeters.

Example: A variable condenser has three rotary plates and four stationary plates. The rotary plates are each of a 6-sq. in. area and are semicircular in shape. The dielectric sheet is of air with a thickness of 0.03 in. Find the capacity of the condenser when the plates are closed or in a maximum capacity position.

Solution:

Using formula (65), first convert inches into centimeters.

$$\begin{aligned} \pi &= 3.1416 \\ 1 \text{ cm.} &= 0.04 \text{ in. (approx.)} \\ 1 \text{ in.} &= 2.54 \text{ cm.} \end{aligned}$$

Hence,

$$C = \frac{2nA}{4\pi d} = \frac{2 \times 3 \times 38.7096}{4 \times 3.1416 \times .0762} = 242 \text{ cm.}$$

To convert centimeters into micro-microfarads multiply by 1.1124 (1 cm. = 1.1124 mmf.).

Answer: 242 cm., or approximately 0.00027 mf. or 270 mmf.

These are all methods of calculating capacity where no measuring instruments are required, only an accurate physical calculation of the geometric structure being necessary. They are given here principally to show what factors affect the capacity of a condenser.

IV. ELEMENTS OF ALTERNATING-CURRENT ENGINEERING

56. Construction of a Sine Wave.—A theoretical operating knowledge of all alternating-current circuits of either a low or high frequency depends primarily upon a fundamental understanding of trigonometry. It is important, therefore, that the reader study these trigonometric principles to obtain a thorough-working knowledge of alternating-current power and radio circuits.

It has been shown that an alternating current, also designated a.c., is a current which changes its direction or flow periodically, starting from zero voltage in one direction, rising to maximum in the opposite direction, and falling again to zero. This condition may be clearly seen by referring to Fig. 58 in which the alternating e.m.f. is indicated by the curve and its phase position at any instant by the moving arm A. This arm is pictured to move in a counterclockwise direction through

360 degrees of a circle which represents one complete cycle of alternating e.m.f. or current. The arm which moves about the circle is called a *vector* or *clock diagram* and represents a definite angular relationship to its horizontal axis, or its vertical component. The angle between the horizontal radius CO and the vector CA (described by the counter-clockwise rotation of the vector) is called the *phase angle* and is expressed by the Greek letter θ (theta). It may be clearly seen that the movement of the vector through 360 degrees of a cycle may serve to indicate the exact relationship of voltage or current with respect to one another or the horizontal axis. There are two types of vectors which may occur in alternating-current problems, namely, *space* and *rotating* vectors.

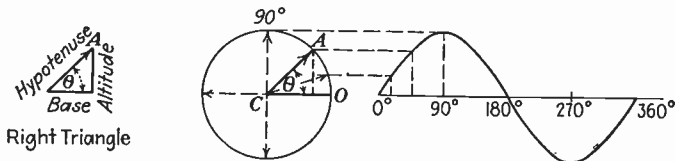


FIG. 58.—A vector or clock diagram of one cycle of alternating current.

Space vectors are those in which there is a fixed direction in space, and rotating vectors are those in which there is a constant magnitude and angular velocity. The latter vectors are also known as *time* vectors.

Alternating currents or e.m.fs. which vary sinusoidally with time are generally expressed by rotating vectors but in such a manner that the rotating vector of fixed magnitude revolves with a constant angular velocity. If the length of the vector represents the maximum value of e.m.f. or current, as the case may be, its projection on a fixed reference axis is the value of the voltage or current at the instant of time considered. That is to say, the number of revolutions per second made by the revolving line is equal to the number of cycles gone through by the voltage or current per second.

Rotating vectors may thus be handled by any of the procedures used in space vectors by merely considering them at some particular instant of time, or they may be treated by methods which are applicable to them alone.

All vectors may be readily analyzed by the application of fundamental trigonometric formulas as applied to the solution of triangles. A knowledge of the solution of the right triangle is absolutely essential since it may be clearly seen that the position of the vector arm with relation to the horizontal axis is always a definite angular expression. Voltage and current relationships may thus be readily comprehended if the various angular functions of trigonometry called *sines*, *cosines*, and *tangents* are clearly understood.

If the instantaneous values of the voltage or current follow one another according to the law of sines, such as in the case of the single-phase alternating-current wave

shape illustrated in Fig. 58, then the maximum voltage or current may be determined by the expression

$$I_{max} \text{ or } E_{max} = \sin \omega t \tag{66}$$

where E_{max} = maximum value of the wave or its *amplitude*.

t = duration of one complete cycle, or the periodic time in seconds.

The abscissa for one cycle is taken as 360 degrees or 2π radians and corresponds to one complete revolution of an armature in a simple 2-pole alternator. In other words, 2 radians correspond to one complete cycle of alternating e.m.f. or current. Thus the abscissa values apply both ways as the dotted lines in Fig. 59 show.

All waves which follow this sine law are called *sinusoidal* or simple-harmonic motions and may serve to express either simple-harmonic voltage or current waves

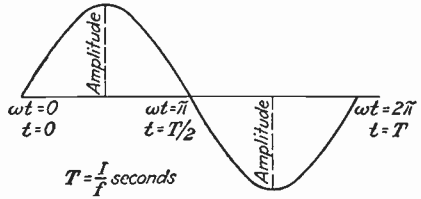


FIG. 59.—Simple sine wave for one cycle.

Referring to the sine wave in Fig. 58, we find that the voltage or current at any instant may then be readily determined by multiplying the sine of the angle by the maximum or peak value of the e.m.f. or current (90-degree value)

Thus,

$$e = E \sin \theta$$

$$i = I \sin \theta$$

where e or i = the instantaneous voltage or current value, and θ = the phase angle ωt

For example, if the peak e.m.f. is assumed to be 120 volts, it will be found that the instantaneous e.m.f. at some part of the cycle, say at 30 degrees, will be $E = \sin \theta$, or $e = 120 \times 0.5$, or 60 volts.

57. Frequency and E.M.F. Formulas.—As has been explained before, a cycle of alternating current is a complete reversal of the current, each cycle consisting of two alternations. The number of cycles taking place per second is referred to as the *frequency* of that particular current.

For example, in an alternator having a frequency of 500 cycles per second there are 1,000 alternations a second. It is preferable to specify the frequency of an alternator, in cycles per second, as for example, a 60-cycle alternator, a 120-cycle alternator, a 500-cycle alternator, and so on.

The following formula shows how the frequency of an alternator may be determined.

$$f = \frac{N}{2} \times \frac{\text{r.p.m.}}{60} \tag{67}$$

where N = number of field poles.

r.p.m. = speed of the armature in revolutions per minute.

Example: A certain alternator has 24 field poles and runs at 3,500 r.p.m. What is its frequency?

$$f = \frac{24}{2} \times \frac{3,500}{60} = 700 \text{ cycles}$$

Example: An alternator is run at 600 r.p.m. and is to give a frequency of 60 cycles. What number of poles is required?

$$N = \frac{2f60}{\text{r.p.m.}} \quad (68)$$

Hence

$$N = \frac{2 \times 60 \times 60}{600} = 12 \text{ poles}$$

It is to be noted that the frequency formula (67) applies only to the revolving-field or revolving-armature types of alternators. It is evident that it could not be applied to an inductor-type alternator. The frequency of an inductor-type alternator is equivalent to the number of inductors which pass a given point per second. For example, the alternator in Fig. 371 has a rotor with 12 teeth. Therefore, to generate a 500-cycle alternating current, it would be necessary that the rotor revolve 41.6 times ($500 \div 12$) per second or approximately 2,500 r.p.m.

Probably the most useful equation used in electrical engineering is the e.m.f. formula. This formula is derived by a consideration of numerous factors. To generate 1 volt it has been determined that it is necessary for a conductor to pass 10^8 lines of force per second. Therefore, the flux density of the field must be represented. The number of conductors connected in series to form the armature must also be considered. Also the number of times each conductor passes a field pole, or, the speed of cutting must be included. In addition, as the flux is dependent upon the area of the pole faces, this area must be considered. All of these factors must be multiplied by a constant K which will depend on certain variable factors, such as arrangement of windings and other design relations known to the designing engineer. The equation may then be stated as

$$E = \phi \times N \times f \times K \quad (69)$$

The constant K combines several figures which become apparent when they are separately analyzed. We know that for every complete revolution the armature conductor passes 2 poles; therefore, $2f$ poles pass a conductor per second. As the conductor must pass 10^8 lines of force (maxwells) per second to generate 1 volt, the average volts per conductor will be $2f \times \phi \div 10^8$. Now, as each armature turn consists of two conductors in series, the volts per turn will be $2 \times 2f \times \phi \div 10^8$ which gives average volts. If E is to equal effective volts, the average volts must be multiplied by 1.11 ($E_{eff} = 1.11 \times E_{av}$). Voltmeters read effective volts.

The constant K then becomes

$$K = \frac{2 \times 2 \times 1.11}{10^8} = \frac{4.44}{10^8} = 4.44 \times 10^{-8}$$

The entire formula for effective E including all of the factors referred to above is written

$$E = \frac{4.44\phi Nf}{10^8} \quad \text{or} \quad 4.44B_m ANf10^{-8} \quad (69a)$$

where B_{max} = maximum normal induction (flux) in maxwells per pole.

A = area of cross section of pole face in square centimeters.

N = number of turns of conductor.

$\phi = B_{max} \times A$.

Note 1: The factor f , which is derived from the frequency formula previously given, involves the number of pairs of field poles and the speed of rotation of the machine when rotating-field or rotating-armature alternators are considered. When the formula is applied to the inductor-type alternator the basic factors are the same although arrived at in a different way, and therefore K also includes these same equivalent factors.

Example: Suppose we have a 4-pole alternator running at a speed of 1,800 r.p.m.; the flux density is 15×10^4 lines; the number of conductors is 300. What is the voltage generated?

Solution: The frequency figures out to 60 cycles and the e.m.f. formula may then be applied:

$$E = \frac{4.44\phi Nf}{10^8} = \frac{4.44 \times 15 \times 10^4 \times 300 \times 60}{10^8} = 119.8, \text{ or approximately } 120 \text{ volts}$$

Note 2: If the alternator in the preceding example was of the inductor type, the same method of figuring could be applied, as f is the figure for frequency and the type of alternator used is of no importance.

58. Effective Voltage and Current.—Since the voltage or current in an alternating-current system is actually of different instantaneous values throughout the time periods of the alternating cycle, and since the cycles follow in rapid sequence per second of time, the actual *effective* voltage or current can only be determined by comparing the alternating-current heating effect with that of direct current.

This is known as the *effective* or *root-mean-square* (*r.m.s.*) value of an alternating current or voltage because, if the instantaneous values of current during a cycle of alternating current are taken, the results squared, and an average value obtained of these squared instantaneous currents, and then the square root of this value derived, the heating effect will be the same as in direct-current circuits, namely, proportional to the square of the current.

Hence the amplitude or peak factor of an alternating wave must be the ratio of its maximum value to its r.m.s. or effective value, or $\sqrt{2} = 1.414$.

Thus

$$E_{eff} \text{ or } E_{r.m.s.} = E_{max} \times 0.707 \quad \text{or} \quad \frac{E_{max}}{1.414} \quad (70)$$

and

$$E_{max} = E_{eff} \times 1.414 \quad \text{or} \quad \frac{E_{eff}}{0.707} \quad (71)$$

Most electrical indicating instruments read the effective-voltage or current values and consequently the peak voltage or current may be readily obtained by multiplying the meter reading by 1.414.

Peak values, particularly for insulators and condensers, are of great importance since the stress to which a dielectric of a condenser is subjected depends upon the maximum value of the alternating-current charging voltage and not upon its r.m.s. or effective value.

59. Effect of Resistance on Phase Relations.—Figure 60 illustrates the phase relationship of the voltage and current through the 360 degrees of a cycle in an alternating-current circuit which does not contain any inductance or capacity but simply ohmic resistance. The vector diagram

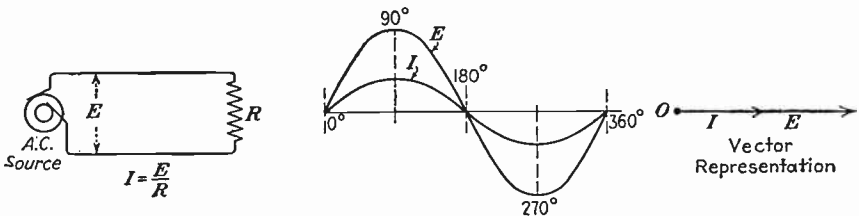


FIG. 60.—Relations in non-reactive circuit.

shows that the voltage and current arms of the rotating vector are in phase, which in other words states that the current and voltage reach their instantaneous periods at the same time and, consequently, are said to be in phase with each other. Hence, $E = IR$ as in direct-current circuits.

60. Effect of Inductance on Phase Relations.—We have seen that an inductance possesses the property of inertia or mass which tends to prevent it from being set into motion and then, once in motion, will tend to prolong its motion. That is to say, the period in which the electrons in a circuit are being accelerated is representative of the work being done in creating a magnetic field around a coil or conductor. Thus, when the accelerating motion is decreased by lowering the source e.m.f. or increasing the resistance of a circuit, the energy stored up in the inductance in the form of a magnetic field is converted into kinetic energy $\frac{1}{2}LI^2$, thereby tending to prolong the movement of the electrons in the conductor. Thus, during the period in which the magnetic field is being expanded around the coil, the current flow will be momentarily opposed resulting in a lagging effect of the current behind the applied voltage. This condition is illustrated in Fig. 61 where the retarding effect acting against the current flow is called the inductive reactance X_L . This is expressed in ohms and

represents the opposition offered to the flow of an alternating current just as does resistance in a direct-current or non-reactive alternating-current circuit. However, since the circuit contains both resistance and reactance (it is assumed that all inductances must have some direct-current resistance), the total opposition must be the result of their combined effects. This is known as the *impedance* of a circuit and is expressed by the letter Z .

If the alternating-current circuit contains pure inductance with absolutely no direct-current resistance, in either the coil or the circuit, the current flow will be 90 degrees behind the source e.m.f., owing to the opposing effect from the counter e.m.f. of self-induction. This condition is illustrated in Fig. 61. The current flow in this circuit may be determined by the ratio $I = E/X_L$, where X_L is the reactive effect in ohms.

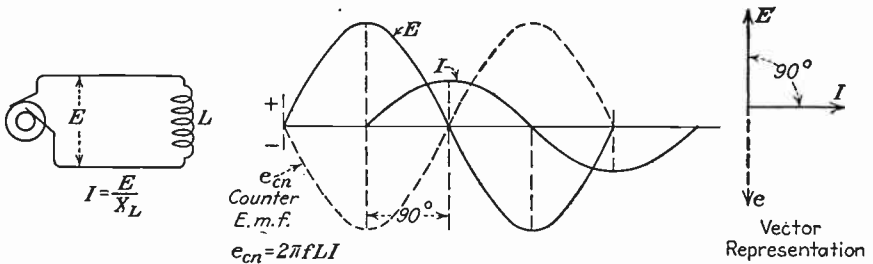


FIG. 61.—Relations in inductively reactive circuit.

The reactance, therefore, is a function of the abscissa for 360 degrees or, 2π radians times the frequency in cycles per second, times the inductance of the circuit, and is generally expressed as ωL or $2\pi fL$. Hence,

$$X_L = \omega L = 2\pi fL \quad \text{ohms.} \tag{72}$$

Now, if we consider that all inductive circuits must contain some resistance and that the reactive and resistive effects are out of phase by 90 degrees, the total effective resistance, or the impedance of the circuit, cannot be obtained as a whole number, or the impedance of the circuit, cannot be obtained as a whole number by adding their respective ohmic values algebraically, but must be obtained by adding their respective values vectorially. In other words, we apply the right triangle to the solution of this problem by using the base as the resistance, or R value, and the altitude as the X_L , or reactive value. The hypotenuse of the right triangle, therefore, will represent the impedance Z of the circuit and is obtained by taking the square root of the sum of the squares [Fig. 62(a)]:

$$Z = \sqrt{R^2 + X_L^2} \quad \text{ohms} \tag{73}$$

Consequently, since the reactance and the resistance effects are added vectorially, it is evident that the voltages across the inductance and the

resistance must also be out of phase, and therefore the effective e.m.f. must also be a vectorial function. Hence, the effective e.m.f. may also be obtained by a right triangle as illustrated in Fig. 62(b):

$$E = \sqrt{E_L^2 + E_R^2} \tag{74}$$

where E_L = the e.m.f. across the inductance.
 E_R = the e.m.f. across the resistance.
 E = the source or the resultant e.m.f.

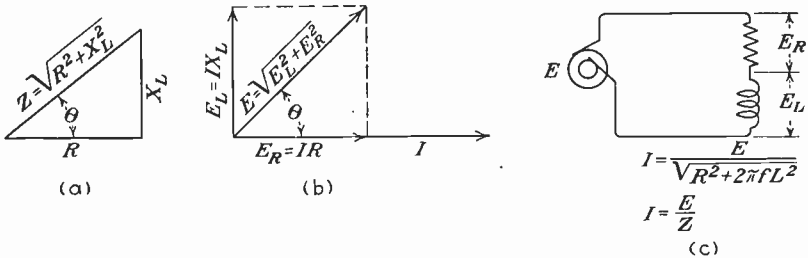


FIG. 62.—Vector representation of inductively reactive circuit containing resistance.

61. Effect of Capacity on Phase Relations.—The condenser, it has been seen, stores up energy in electrostatic form owing to the displacement of the electrons in the dielectric. During the period of this displacement, the condenser is building up a potential difference which eventually equals the source e.m.f. After this, no more electrons will be moved throughout the circuit. Now, if the applied potential is of an alternating-current characteristic, the condenser will charge from 0 to 90 and 180 to 270 degrees of the cycle and it will discharge through the circuit from 90 to 180 and 270 to 360 degrees of the cycle. That is to say the condenser is offering the maximum opposition to the source e.m.f. at the 90- and 270-degree periods and the least opposition at the beginning of the half-cycle periods. If, therefore, a condenser is permanently connected across an alternating-current source, a continuous current will flow in the condenser circuit owing to the dielectric displacements for each reversal of source potential. The amount of current flowing will depend upon the e.m.f., the capacity, and the number of reversals per second (the frequency of the source e.m.f.). Hence, the greater the frequency applied to the condenser plates, the greater will be the number of displacements in the dielectric and the greater will be the flow of current in the circuit. In other words, the reactance that the condenser offers to a higher frequency will be less than at lower frequencies, and, consequently, the reactance X_c due to the condenser is said to vary inversely with the frequency.

Obviously, since the condenser reactance varies inversely with the frequency, the formula for determining this would be expressed:

$$X_c = \frac{1}{\omega C} = \frac{1}{2\pi f C} \quad \text{ohms} \quad (75)$$

where C = the capacity in farads.

However, since all capacities are generally in micro-units or microfarads, the formula may be more conveniently expressed

$$X_c = \frac{1,000,000}{2\pi f C} \quad \text{ohms}$$

where C is in microfarads.

Figure 63(a), (b), and (c) illustrates the *leading* effect of the current ahead of the e.m.f. at the various points of the cycle. If the condenser,

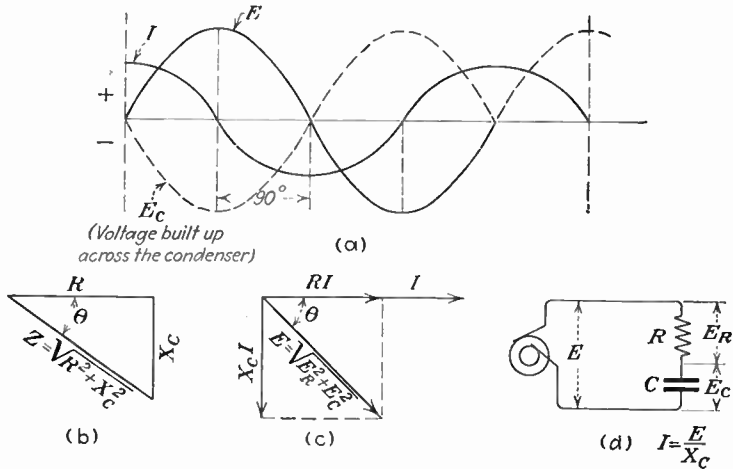


FIG. 63.—Vector and phase representation of capacitively reactive circuit containing resistance.

or the circuit in which it is connected, contains no direct-current resistance, the current will lead by 90 degrees, but since this condition is practically impossible the angle of lead will be somewhat less than 90 degrees depending upon the amount of resistance introduced. If, however, the resistance is negligible, then the current flow in the circuit can be readily computed from the ratio

$$I = \frac{E}{X_c} \quad (76)$$

If, on the other hand, resistance is to be included in the calculation, the expression for impedance must again be used, namely,

$$I = \frac{E}{Z} \quad (77)$$

Thus it can readily be seen that in all forms of reactive circuits, either inductive or capacitive, which contain resistance, the current and voltage

relationships in terms of *lead* or *lag* may be determined by the phase angle θ , expressed trigonometrically, $\cos \theta$. In other words, it is possible to determine the exact number of degrees between 0 and 90 degrees that a voltage or current will lead or lag, as the case may be, owing to the resistance in the circuit. Expressed mathematically, the angular difference may be obtained by any of the following relations:

$$\cos \theta = \frac{R}{Z} \quad (78)$$

$$\sin \theta = \frac{X}{Z} \quad (79)$$

$$\tan \theta = \frac{X}{R} \quad (80)$$

62. Power Factor.—In alternating-current engineering the equation (78) is known as the *power factor* of a circuit and is used to indicate the exact ratio in percentage between any two sinusoidal waves of either current, voltage, or both, relative to each other.

Example: Assume the source voltage in Fig. 63(d) to be 120 volts at 60 cycles; the resistance R , 15 ohms; and the condenser C , 25 mf. What is the angle of lead in percentage and in degrees?

Solution:

$$X_C = \frac{1,000,000}{2\pi fc} = 106^\omega$$

$$Z = \sqrt{R^2 + X_C^2} = \sqrt{15^2 + 106^2}$$

$$Z = \sqrt{225 + 11,236} = 107$$

Answer:

$$\text{Power factor} = \frac{R}{Z} = \frac{15}{107} = 0.14 \text{ per cent}$$

The angle whose cosine = 0.14 is 82 degrees (see any table of natural sines and cosines).

Hence the current in this example leads the voltage by 82 degrees.

63. Ohm's law for alternating-current circuits containing reactive and resistive properties may thus be expressed by the ratio,

$$I = \frac{E}{\sqrt{X_{L \text{ or } C}^2 + R^2}} = I = \frac{E}{Z} \quad (77)$$

$$Z = \frac{E}{I} \quad (81)$$

$$E = IZ \quad (82)$$

Example: The circuit in Fig. 62(c) contains a direct-current resistance of 4 ohms and an inductance of 680 mh. The alternating-current effective e.m.f. is 500 volts at a frequency of 120 cycles. Find the value of the current flowing in the circuit.

Solution:

$$X_L = 2\pi fL = 2 \times 3.14 \times 120 \times 0.68 = 512 + \text{ohms}$$

$$R = 4 \text{ ohms}$$

$$Z = \sqrt{R^2 + X_L^2} = \sqrt{16 + 262,308} = 512 \text{ ohms}$$

Answer:

$$I = \frac{E}{Z} = \frac{500}{512} = 0.97 \text{ amp.}$$

It can be clearly seen from this example that the current flow will be practically dependent upon $I = E/X_L$, if the direct-current resistance is very low. For practical purposes this ratio may be sufficient, but in cases where rigid accuracy is desired the impedance function must be maintained.

64. Phase Relations in Series-resonant Circuits.—A circuit which contains inductance and capacitance in series is generally known as a

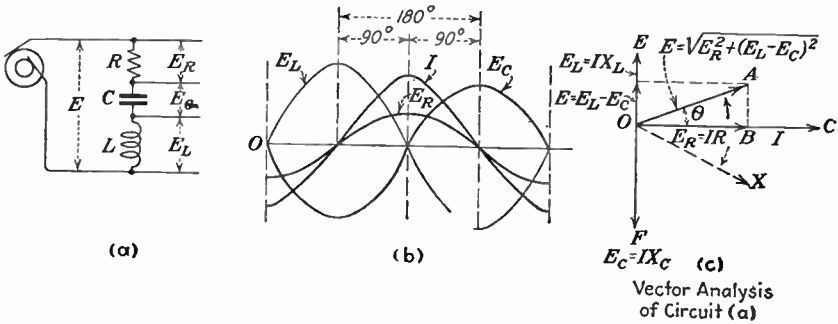


Fig. 64.—Vector and phase representation of series-resonant circuit.

series-resonant circuit. This is true regardless of whether there is any direct-current resistance present in the circuit or not. A series-resonant circuit, together with a vector analysis of its current and voltage relations, is illustrated in Fig. 64(a), (b), and (c).

We have seen that the voltage across an inductance and a capacity in a single inductive or capacitive circuit will always be out of phase with the current flow in the circuit. The voltage and current across a resistance, however, will be in phase. Now, if an inductance, a capacity, and a resistance are connected in series with each other the phase relations between each component part of L , C , and R will be as illustrated in Fig. 64(b) and (c).

The current flow in a series circuit is the same at any portion of the circuit and may be conveniently expressed by the horizontal line OC in Fig. 64(c). The e.m.f. across the resistance ($E_R = IR$) is measured by the vector OB which coincides with the direction of OC . That is to say, they both move together, since the voltage and current in a resistive circuit are in phase. The e.m.f. of self-induction built up across the inductance E_L is represented by the vector arm OE and is shown at right angles to the current to indicate that the current due to the inductance is lagging by 90 degrees behind the e.m.f. Similarly, the e.m.f. built up across the condenser E_C is represented by the vector arm OF , also at

right angles to the current but in the opposite direction, to show that the current due to the condenser is leading the e.m.f. by 90 degrees.

Assuming that the voltage across the inductance E_L is of the same value as the voltage across the condenser E_C , the two e.m.f.s. would act in opposite directions to each other (oppose) owing to the fact that they are 180 degrees apart in their phase relationship. If this were the case, then their combined effects would be equal to their arithmetical difference $E_L - E_C$. This condition is represented by the vector BA which represents the combined e.m.f.s. due to the inductive and capacity effects, and OA which represents the applied or source e.m.f., thus completing the right-triangle expression of e.m.f. in a series-resonant circuit.

It will be noted in Fig. 64(c) that the vector arm OA is assumed to be rotating in a counterclockwise direction. This merely serves to indicate that the inductance E_L in the circuit predominated in its reactive effect relative to the condenser, and, therefore, the arm is pictured as moving in the direction OE . Hence the phase-angle sign θ is shown above the horizontal axis OB in the direction of OE to indicate that the current flow in the entire circuit is lagging behind the e.m.f. by a certain number of degrees or $\cos \theta$. If, on the other hand, it is assumed that the capacitive reactive effect E_C predominates, then the arm OA would be pictured to act in the direction of OF , as illustrated by the imaginary arm OX . Under this condition, the phase angle θ would indicate the number of degrees that the current leads the e.m.f. in the circuit.

Hence the triangle OBA may be expressed mathematically

$$E = I\sqrt{R^2 + (X_L - X_C)^2} = IZ$$

Thus,

$$Z = \sqrt{R^2 + \left(2\pi fL - \frac{1}{2\pi fC}\right)^2} = \sqrt{R^2 + X^2}$$

Therefore,

$$I = \frac{E}{\sqrt{R^2 + X^2}} = \frac{E}{Z}$$

A practical example may be applied to Fig. 64(a) in which the source frequency f is 120 cycles; the resistance R , 20 ohms; the inductance L , 2 henrys; and the condenser C , 10 mf. A current of 3 amp. is flowing in the circuit.

Example: Calculate the following:

- (a) The e.m.f. across the component parts R , C , and L .
- (b) The applied e.m.f. E .
- (c) The circuit impedance Z .
- (d) The power-factor cosine θ .

Solution:

$$X_L = 2\pi \times 120 \times 2 = 1,507 \text{ ohms}$$

$$X_C = \frac{1,000,000}{2\pi \times 120 \times 10} = 133 \text{ ohms}$$

$$(a) \quad E_R = I \times R = 3 \times 20 = 60 \text{ volts}$$

$$E_c = I \times X_c = 3 \times 133 = 399 \text{ volts}$$

$$E_L = I \times X_L = 3 \times 1,507 = 4,521 \text{ volts}$$

(b) $E = \sqrt{E_k^2 + ((E_L - E_c)^2)}$

$$E = \sqrt{60^2 + (4,521 - 399)^2} = 4,122.4 \text{ volts}$$

(c) $Z = \sqrt{R^2 + (X_L - X_C)^2}$

$$= \sqrt{20^2 + (1,507 - 133)^2} = 1,374 \text{ ohms}$$

(d) Power factor = $\cos \theta = \frac{R}{Z} = \frac{20}{1,374} = 0.0145$ or 89 degrees

or

$$\text{Power factor} = \cos \theta = \frac{P}{E \times I} = \frac{\text{true watts}}{\text{apparent watts}}$$

Hence the inductance in this circuit predominates, and the current lags behind the e.m.f. by a phase difference of 89 degrees.

65. Phase Relations in Parallel-resonant Circuits.—Unlike the series-resonant circuit the current flow is not the same in any portion of these circuits but is dependent upon the current flow in each of the resistive and reactive branches. The current flow, of course, also depends

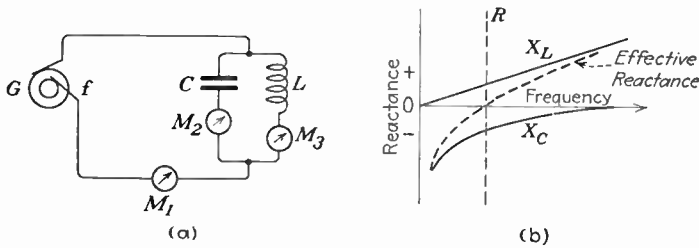


FIG. 65.—Parallel resonant circuit.

upon the applied or source e.m.f. and is the same across each parallel branch as in the simple parallel-resistance circuit.

Parallel-resonant circuits are perhaps the most commonly used in transmitting and receiving circuits. An understanding of their functioning is of great importance for a thorough understanding of radio circuits.

Let us assume the generator *G* in Fig. 65(a) to be delivering a constant voltage at the zero frequency, and for the sake of convenience this voltage will be assumed to remain constant at all times regardless of the frequency. Now picture the frequency of the generator rising very slowly from zero and continuing to infinity. When the frequency is relatively low, the current flow through the condenser will be at a minimum and that flowing through the inductance at a maximum. These conditions will be indicated by a relatively greater reading through the meter *M*₃ in the inductively reactive circuit as compared to the reading of the meter *M*₂ in the capacitively reactive circuit. This result can be clearly seen if we recall that the reactance due to an inductance *X*_L varies directly with the

frequency, and the reactance due to a condenser X_c varies inversely with the frequency.

Hence, as the frequency of the generator is gradually increasing, the current through the branch M_3 will be decreasing, and that through the branch M_2 will be increasing until the meters read exactly the same. If the frequency is continued to infinity, these conditions will intensify; that is, the current through the inductance will become smaller and smaller and that through the condenser will become greater and greater. Let us assume, however, that the alternator frequency is held constantly at the point in which both meters M_2 and M_3 read alike. We can clearly see that at this point both the reactances X_L and X_C are equal, as shown in Fig. 65(b), and, since their phase relations are opposite, the effective reactance will be zero. This is called the *resonant condition* (dotted line R in diagram) in a parallel-resonant circuit and indicates that

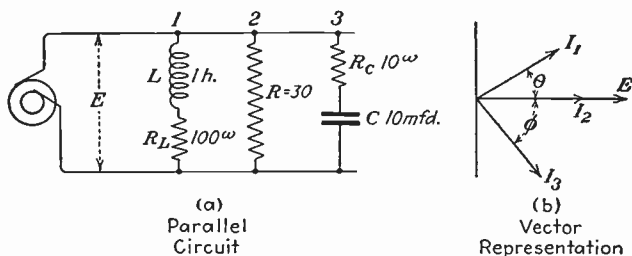


FIG. 66.—Parallel impedance branches.

a maximum value of current will flow in the parallel mesh L and C , limited only by the direct-current resistance in the coil and the condenser.

It is very important to note that at the resonant frequency the meter M_1 will read practically zero and *not* the sum of the branch currents. This is due to the fact that the *electrons circulating in the mesh circuit* are opposed by the electrons in the feed circuit, that is to say, as the electrons in the mesh circuit are endeavoring to enter into the supply line, the electrons in the supply line are also trying to enter into the parallel mesh, and, since both circuits are varying in phase at the same frequency, the net result of the electron flowing through the meter M_1 will be the vectorial sum of the two branches, which at the absolute resonant frequency is zero.

66. Parallel-circuit Analysis.—Figure 66(a) illustrates a typical parallel-resonant circuit with a vector representation [Fig. 66(b)] of the current flow in the various branches. The sign θ (theta) is used in parallel-circuit vectors to represent the phase angle due to the inductance, and the sign ϕ (phi) to denote the phase angle due to the capacitance. Both of these signs may be interchanged at will since they only express opposite conditions or vectorial directions.

The analysis of a parallel circuit may be simplified considerably if each of the branches is taken separately as a simple series circuit especially since the e.m.f. E across each branch will be the same.

Then by applying Ohm's law to each branch the individual branch currents may be determined:

Step 1:

$$I_L = \frac{E}{Z} \quad (77)$$

$$I_C = \frac{E}{Z} \quad \text{or} \quad I_C = 2\pi fCE \quad (83)$$

$$I_R = \frac{E}{R} \quad (5)$$

If either of the reactive circuits is assumed to have negligible resistance, the denominator in I_L and I_C should be changed from Z to read X_L and X_C , respectively.

Now, since the current flow through the inductive and the capacitive branches are opposite angular expressions, their net result must be a vectorial quantity $I_L - I_C$. Hence, if the resistance component is to be included, the total current flow in the circuit would be

Step 2:

$$I_{\text{total}} = \sqrt{I_R^2 + (I_L - I_C)^2} \quad (84)$$

The entire circuit impedance, after the total current flow has been found, may then be obtained from the ratio

Step 3:

$$Z = \frac{1}{Y} = \frac{1}{\sqrt{R^2 + \left(\frac{1}{2\pi fL} - 2\pi fC\right)^2}} \quad (85)$$

The denominator in this equation is called the *admittance* Y and is the reciprocal of the impedance.

It will be remembered that the impedance of a series-resonant circuit is obtained by the expressions $2\pi fL$ and $1/2\pi fC$; therefore, in parallel-resonant circuits the conditions must be changed to $1/2\pi fL$ and $2\pi fC$. These two quantities are called the *susceptances* of the inductance and the condenser respectively. Now, just as conductance is generally expressed as the reciprocal of the resistance, susceptance is the reciprocal of the reactance, and admittance the reciprocal of the impedance. Although this relationship generally holds true, it is important to remember that the susceptance is not a reciprocal of the reactance unless the resistance is zero. The relation is true of the conductance in reactive circuits unless the reactance is zero. Admittance, on the other hand, is always the reciprocal of the impedance since it includes the combined functions in the ratio.

The susceptance is expressed by b and the conductance by g .

When there are a number of reactive components in parallel, their resulting conductance may be computed by adding the individual conductances;

$$G = g_1 + g_2 + g_3 \text{ etc.} \quad (86)$$

Similarly, the total susceptances may be obtained by adding the susceptances

$$B = b_1 + b_2 + b_3 \text{ etc.} \quad (87)$$

When analyzing parallel circuits containing both positive and negative reactances (inductive reactance is positive and capacitive reactance is negative), the susceptances must also be considered with $+$ and $-$ signs. Consequently, when various branch susceptances are to be added together, the addition must be algebraic in order to keep all results accurate with relation to phase changes.

Thus

$$G = \frac{R}{Z^2} \text{ mhos} \quad (88)$$

$$B = \frac{X}{Z^2} \text{ mhos} \quad (89)$$

Example: Assuming the applied frequency in Fig. 66 (a) to be 60 cycles at a source e.m.f. of 220 volts. The inductance has a value of 1 henry with a direct-current resistance of 100 ohms. The condenser has a value of 10 mf. and a series resistance R_c of 10 ohms. The non-reactive resistance R has a value of 30 ohms. Find the following:

- The individual component impedances Z_1 , Z_2 , and Z_3 .
- The individual component currents I_1 , I_2 , I_3 .
- The individual component conductances g_1 , g_2 , g_3 .
- The individual component susceptances b_1 , b_2 , b_3 .
- The admittance Y .
- The impedance Z .

Solution:

- (a) Branch reactances and impedances:

$$Z_1 \sqrt{R^2 + X_L^2} = \sqrt{10,000 + 141,978.24} = 390^\omega \quad (\text{approx.})$$

$$Z_2 = 30 \text{ ohms}$$

$$Z_3 = \sqrt{R^2 + \left(\frac{1,000,000}{2\pi fC}\right)^2} = \sqrt{100 + 70,225} = 265^\omega \quad (\text{approx.})$$

- (b) Branch currents:

$$I_1 = \frac{E}{Z} = \frac{220}{390} = 0.56 \text{ amp.}$$

$$I_2 = \frac{E}{R} = \frac{220}{30} = 7.3 \text{ amp.}$$

$$I_3 = \frac{E}{Z} = \frac{220}{265} = 0.83 \text{ amp.}$$

- (c) Branch conductances:

$$g_1 = \frac{R}{Z_1^2} = \frac{100}{152,100} = 0.0007 \text{ mho}$$

$$g_2 = \frac{R}{R_2^2} = \frac{30}{900} = 0.033 \text{ mho}$$

$$g_3 = \frac{R}{Z_3^2} = \frac{10}{70,225} = 0.00014 \text{ mho}$$

(d) Branch susceptances:

$$b_1 = \frac{X_1}{Z_1^2} = \frac{376.8}{152,100} = 0.0024 \text{ mho}$$

$$b_2 = 0$$

$$b_3 = \frac{X_3}{Z_3^2} = \frac{265}{70,225} = 0.0038 \text{ mho}$$

Hence the total susceptance:

$$B = b_1 + b_3 = (0.0024) + (0.0038) = 0.0062 \text{ mho.}$$

Note: b_2 is zero because the current in this branch is in phase with E .

(e) Admittance:

$$Y = \sqrt{G^2 + B^2} = \sqrt{0.0338^2 + 0.0062^2} = 0.034 \text{ mhos (approx.)}$$

(f) Impedance:

$$Z = \frac{1}{Y} = \frac{1}{0.034} = 30 \text{ ohms (approx.)}$$

Proof: When the applied E is given, we use the ratio for parallel circuits:

$$Z = \frac{E}{\sqrt{I_R^2 + (I_L - I_C)^2}}$$

Hence, in the example,

$$Z = \frac{E}{I} = \frac{220}{7.3} = 30 \text{ ohms (approx.)}$$

If it is assumed that the resistances R_L and R_c are negligible, then the impedance may be more easily determined by the ratio

$$Z = \frac{1}{\sqrt{\left(\frac{1}{R}\right)^2 + \left(\frac{1}{X_C} - \frac{1}{X_L}\right)^2}}$$

Thus, in the example,

$$Z = \frac{1}{\sqrt{\left(\frac{1}{30}\right)^2 + \left(\frac{1}{265} - \frac{1}{377}\right)^2}} = \frac{1}{0.033} = 30 \text{ ohms (approx.)}$$

67. Resonance.—Electrical resonance is that condition encountered in series- or parallel-resonant circuits when the inductive and capacitive reactances are equal and balance each other so that their reactive effect upon the circuit will be zero. Since an inductive reactance introduces a lagging effect, and a capacitive reactance a leading effect, on the current with respect to the e.m.f., the resultant current flow will be dependent only upon the direct-current resistance in the circuit. Therefore, since the current and voltage across the direct-current resistance are always in

phase with each other, the current flow in a series-resonant circuit will be $I = E/R$.

It may thus be readily seen that if the impedance in a series-resonant circuit is dependent upon $X_L = X_C$, then, by substituting and equating $\omega L = 1/\omega C$ and solving for the frequency, it would be possible to determine the resonant frequency for any value of L and C in a series-resonant or parallel-resonant circuit by the formula

$$f = \frac{1}{2\pi\sqrt{LC}} \text{ cycles,} \quad \left(L = \frac{1}{4\pi^2 f^2 C} \text{ henrys,} \quad C = \frac{1}{4\pi^2 f^2 L} \text{ farads} \right) \quad (90)$$

where L and C are expressed in henrys and farads, respectively.

Since all capacities are in microfarads, and inductance in either micro-, milli- or unit henrys, the frequency formula may be simplified considerably by using the proper numerator to fit the desired condition. For example, if it is desired to determine the resonant frequency of a series-resonant circuit in which both L and C are in microhenrys and microfarads, respectively, then the formula may be expressed:

$$f = \frac{1,000,000}{2\pi\sqrt{LC}} \text{ cycles,} \quad \left(L = \frac{1}{4\pi^2 f^2 C} \text{ henrys,} \quad C = \frac{1}{4\pi^2 f^2 L} \text{ farads} \right) \quad (a)$$

If, on the other hand, the inductance and capacity are in henrys and microfarads, respectively, then the formula may be expressed

$$f = \frac{1,000}{2\pi\sqrt{LC}} \text{ cycles} \quad (b)$$

Equation (90) is the general formula for determining resonant frequency, while equation (a) is generally used for high-frequency radio circuits and equation (b) for low-frequency and filtering circuits.

Example: Assume L and C in Fig. 64(a) to have a value of 15 henrys of inductance and 4 mf. of capacity. At what frequency is this circuit resonant?

Solution:

$$f = \frac{1,000}{2\pi\sqrt{LC}} = \frac{1,000}{6.28\sqrt{60}} = \frac{1,000}{48.67} = 21 \text{ cycles}$$

It will be noted that the circuit resistance is not introduced into this formula. In general practice, R need not be considered if the ratio $4 \times L/C$ is great compared with the value of R^2 .

68. The J Operator. *a. Notation.*—In practical electrical engineering where a number of impedances must be, or are likely to be mathematically treated together to solve a problem, notation is simplified by labeling all the reactive components of the impedances j (either $+j$ or $-j$), to indicate that these components lie along the vertical axis of the vector and are inductive or capacitive, respectively. No numerical value need be

attached to j under certain conditions, and it is then only used to indicate that the numerical quantity which follows is either inductive or capacitive reactance. Under other conditions, j has a numerical value, as will be brought out in the explanation which follows.

Among other special notations applied in electrical engineering is the symbol enclosed in bars to indicate *absolute value of*. Thus $|Z|$ denotes the absolute value or magnitude (vectorial length) of impedance. A line (or dot) is used over the symbol to indicate that a *complete* vector is described. Thus \bar{Z} denotes a magnitude composed of two components, R and X . If vector notation is obvious, the line or dot is dispensed with.

b. Vectors.—We have learned from our consideration of a single impedance and the e.m.f. across it that both of the quantities (Z and E_A) may be resolved into an active and a reactive component in quadrature; that is, R and X , or E_R and E_X , at right angles to each other. In engineering problems in which we have to deal with a number of impedances, e.m.fs., or currents in different phase, the solution of the problem generally becomes a relatively simple matter once these components are known. We may take as a simple illustration the circuit shown in Fig. 67(a) in which there are two positive impedances Z_1 and Z_2 connected in series. Their values are respectively 5 and 5.5 ohms, and their phase angles are given as $\phi_1 = 37$ degrees and $\phi_2 = 69$ degrees. If the active (R) and reactive (X) components of the two impedances were known, we could readily obtain the equivalent or resultant impedance of the two in series by adding these components in accordance with

$$|Z| = \sqrt{(R_1 + R_2)^2 + (X_1 + X_2)^2}$$

But, with only the impedance and phase values given, we should first have to solve for the R and X components before we could make use of the above expression.

An equation of the type just mentioned might be solved trigonometrically, since we know $R_1 = Z \cos \phi_1$ and $X_1 = Z_1 \sin \phi_1$, and similarly in the case of R_2 and X_2 . It is obvious that this method would become quite involved, if we had a large number of impedances to deal with. For this reason, and in order to illustrate the so-called *method of complex quantities*, we shall solve the above problem by simple vector addition. It should be pointed out that the impedance is a so-called *complex quantity*. It is, however, referred to as a *vector impedance* since such quantities are treated as vectors when performing the operation of addition, subtraction, multiplication, or division.

It has already been pointed out that a vector is a quantity having not only magnitude but also direction, and that the phase angle between Z and R , which apparently gives Z a direction, really refers to the effect which the impedance has upon the phase of the current.

We may, therefore, represent the vector impedances of our problem as a phase vector as shown in Fig. 67(b). The vectors are 5.0 and 5.5 units in length, respectively, to correspond with the impedance values, and each makes the required angle with the horizontal axis chosen for reference. It is obvious that the R components of \bar{Z}_1 and \bar{Z}_2 lie along this axis since the phase angle ϕ_1 is, as we know, the angle between \bar{Z}_1 and R_1 , and ϕ_2 that between \bar{Z}_2 and R_2 .

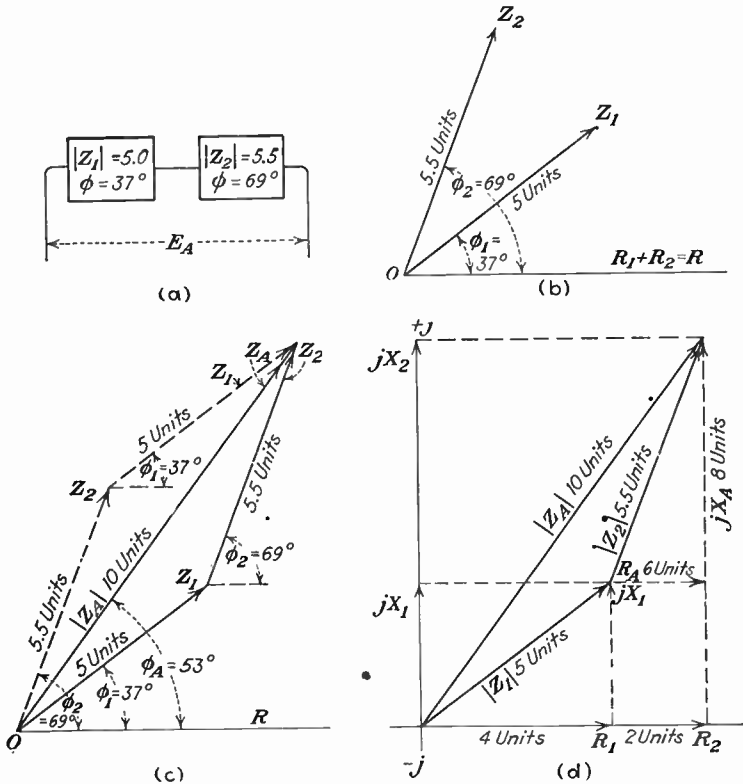


FIG. 67.—Vector representation of series impedances.

To obtain the vector sum of \bar{Z}_1 and \bar{Z}_2 we proceed as shown in Fig. 67(c). The two vectors are placed end to end without changing their original direction. The resultant vector \bar{Z}_A which is drawn from the point O of the first vector to the end point of the last vector then represents the vector sum of the components \bar{Z}_1 and \bar{Z}_2 . The order in which the component vectors (\bar{Z}_1 and \bar{Z}_2) are joined is of no consequence, for, so long as each component is properly directed, the resultant obtained will be the same. This fact is illustrated by means of the dash lines in Fig. 67(c) showing the components taken in reverse order.

The vector \bar{Z}_A gives us the total impedance and phase of $\bar{Z}_1 + \bar{Z}_2$. Its length represents the magnitude or value of the resultant impedance; that is, $|Z_A| = 10$ units (ohms), and the angle which it makes with the reference axis represents the resultant phase angle, that is, $\phi_A = 53$ degrees. The relations shown in Fig. 67(b) and (c) also apply to the voltages involved but only in a relative manner.

The vector sum of any number of component vectors may be obtained in similar manner to that described above. This method of solution is therefore very useful whenever a number of vector quantities, such as impedances or voltages, which are in series, are to be added without recourse to their active and reactive components.

When we wish to show that a vector sum of a number of vectors, or complex quantities, is to be taken we indicate that fact in vector notation. Thus, in general, $\bar{A} = \bar{B} + \bar{C} + \bar{D}$, etc., means that these quantities are to be added vectorially. In the case of alternating-current quantities, such as impedances in series, for example, we should, therefore, write $\bar{Z}_A = \bar{Z}_1 + \bar{Z}_2 + \bar{Z}_3 + \dots$, etc. However for the purpose of brevity, vector notation is often omitted in such problems since it is obvious that none other than a vector sum could be intended. Thus for example, $Z_A = Z_1 + Z_2$ etc. or $E_A = E_1 + E_2$, etc., could obviously not mean the arithmetical sum of these quantities since we know that, if we are dealing with alternating-current values, these quantities are all vectors and cannot, therefore, be added arithmetically. In our last problem, for example, the arithmetical sum of the impedance values would have been 10.5 ohms instead of 10 ohms. In other words $|Z_A| \neq |Z_1| + |Z_2|$. It must be understood, therefore, that the notation $Z_A = Z_1 + Z_2$ cannot refer to an arithmetic sum of values but must refer to a vector sum.

Referring again to Figs. 67(b) and (c), it should be noted that, when an impedance is represented as a vector and definitely referred to some reference axis by means of its phase angle, the vector virtually represents the hypotenuse of an impedance triangle, the R component of which lies along this reference axis. This as well as the fact that the X component is always at right angles to R makes it possible to indicate both the R and X components in the case of each vector, as shown by the dash lines in Fig. 67(d).

The values of these components are found to be as follows:

Vector	Value of R and X Components	
\bar{Z}_1	$R_1 = 4$	$X_1 = 3$
\bar{Z}_2	$R_2 = 2$	$X_2 = 5$
—	—	—
\bar{Z}_A	$R_A = 6$	$X_A = 8$

It is evident that the sum of the R and X components of the component vectors \bar{Z}_1 and \bar{Z}_2 are equal, respectively, to the R and X components of the resultant vector \bar{Z}_A ; that is, $R_1 + R_2 = R_A$ and $X_1 + X_2 = X_A$. Our previous assumption that the resultant of any number of component

vectors expresses the vector sum of these components is verified by the fact that the vector sum of the R and X components of any vector represents the vector itself, that is, each vector may be considered as the resultant of its R and X components.

It is also evident in Fig. 67(d) that any vector is just as definitely determined by its R and X components as by its magnitude (length) and phase angle. This is obvious when we consider that the vector is really the resultant or vector sum of its components. For example, the magnitude and phase of \bar{Z}_1 are determined when R_1 and X_1 are given as

$$|Z_1| = \sqrt{R_1^2 + X_1^2} \quad \text{and} \quad \tan \phi_1 = \frac{|X_1|}{R_1}$$

Since the angle which the vector Z_1 makes with the reference axis is the phase angle ϕ_1 , we know that R_1 must lie along that axis. Since the only requirement in the case of X_1 is that it shall be at right angles to R_1 , we may place it so anywhere on the reference axis. It may be placed in the position indicated by the vertical dash line at the end point of R_1 , or in a position jX_1 along the vertical coordinate axis marked $(-j)(+j)$. We may designate this axis as the one along which all X components are to be referred [see Fig. 67(d)].

c. Complex Quantities: The J Operator.—This conventional, geometrical method of dealing with vector quantities, in which the vectors are definitely referred to a coordinate system of axes and resolved into two components, one of which is along and the other at right angles to a reference axis, is, with the exception of minor details, identical with the so-called method of complex quantities. In the latter method the horizontal axis, which is usually also chosen for reference, is called the *axis of reals*. The components referred to this axis are called the real components, or “reals.” The axis at right angles [$(-j)$, $(+j)$] is called the axis of “imaginaries.” The components referred to it are called imaginary components, or “imaginaries.” They are, of course, not really imaginary components, any more than are the components referred to the so-called axis of reals. The imaginary components are prefixed with $+j$ when they are to be measured along the positive $(+j)$ axis, and with $-j$ when they are to be measured along the negative $(-j)$ axis.

Assuming that our vector quantities in Fig. 67(d) are referred to such a system of coordinate axes, we may write these quantities in *complex* instead of vector form. Under these conditions impedance is expressed by the equation

$$\bar{Z} = R + jX \quad (91)$$

Complex quantities are then written:

COMPONENTS

	Reals	Imaginarys	
$Z_1 =$	R_1	$+ jX_1$	$= 4 + j3$
$Z_2 =$	R_2	$+ jX_2$	$= 2 + j5$
<hr style="width: 100%;"/>	<hr style="width: 100%;"/>	<hr style="width: 100%;"/>	<hr style="width: 100%;"/>
$Z_A =$	R_A	$+ jX_A$	$= 6 + j8$

Thus our problem of two impedances in series would be represented by writing $\bar{Z}_A = \bar{Z}_1 + \bar{Z}_2 = (R_1 + jX_1) + (R_2 + jX_2)$. To solve this complex expression, the R and X components must be known. The reals and imaginarys are then added separately, thus:

$$\begin{array}{r} Z_1 = R_1 + jX_1 = 4 + j3 \\ Z_2 = R_2 + jX_2 = 2 + j5 \\ \hline Z_A = R_A + jX_A = 6 + j8 \end{array}$$

The magnitude of the resultant impedance, $\bar{Z}_A = 6 + j8$, is given by

$$|Z_A| = \sqrt{R_A^2 + X_A^2} = \sqrt{6^2 + 8^2} = 10 \text{ ohms}$$

and its phase angle is

$$\tan \phi_A = \frac{8}{6} = 1.33$$

or

$$\phi_A = 53 \text{ degrees}$$

If, instead of the impedances, the voltage drops across them had been given in our problem they would have been represented in the same way as that shown in Fig. 67(b), (c), and (d). If the phase angles ϕ_1 and ϕ_2 were the same, the diagram of voltages would be geometrically similar; in other words, the impedance and voltage components are directly related through the current which is common to all of the circuit elements when they are in series. This similarity was pointed out in connection with Fig. 64(c). The expressions for the vector voltages written as complex quantities, therefore, resemble those for the impedances. Thus,

$$\begin{aligned} E_1 &= E_{R_1} + jE_{x_1} \\ E_2 &= E_{R_2} + jE_{x_2} \end{aligned}$$

where

$$\begin{array}{ll} E_{R_1} = \bar{I}R_1 & \text{and} \quad jE_{x_1} = jIX_1 \\ E_{R_2} = \bar{I}R_2 & \text{and} \quad jE_{x_2} = jIX_2 \end{array}$$

and their sum

$$\bar{E}_A = \bar{E}_1 + \bar{E}_2 = (E_{R_1} + E_{R_2}) + j(E_{x_1} + E_{x_2}); = (\bar{I}\bar{Z}_A)$$

The real and imaginary components are added in the same way in the case of any number of vector voltages in series.

To illustrate the numerical solution in a simple manner let us assume that the diagram of voltages is geometrically similar to that shown in

Fig. 67(d) and that the voltage drop $\bar{E}_1 = 50$ volts and $\bar{E}_2 = 55$ volts. Since $\bar{E}_1 = I\bar{Z}_1$ and $|Z_1| = 5$ ohms, the current $I = 10$ amp.

Then,

$$\begin{aligned} E_{R_1} &= 40; & jE_{x_1} &= j30 \text{ and } \bar{E}_1 = 40 + j30 \\ E_{R_2} &= 20; & jE_{x_2} &= j50 \text{ and } \bar{E}_2 = 20 + j50 \end{aligned}$$

The applied voltage, which is the sum of those components, is

$$\bar{E}_A = \bar{E}_1 + \bar{E}_2 = 60 + j80,$$

and its value $|E_A| = \sqrt{(60)^2 + (80)^2} = 100$ volts and

$$\tan \phi E_A = 80/60 = 1.33,$$

that is, $\phi E_A = 53$ degrees, which is the value of the phase angle of Z_A .

It has been pointed out that the symbol j refers to the component which is at right angles to the axis of reals. Since this component is regarded as having been operated upon by j , as pointed out before, and thereby rotated 90 degrees with respect to the axis of reals, j is called an *operator*. The operation indicated by plus j , or simply j , is a positive or counterclockwise rotation of 90 degrees as we have seen. Minus j ($-j$) indicates a negative or clockwise rotation of 90 degrees, that is, the imaginary component makes an angle of -90 degrees with respect to the axis of reals. A component rotated through 180 degrees, positively or negatively, would be indicated by $j \times j$ or $(-j) \times (-j) = j^2$. After such operation the component will obviously again lie along the axis of reals, but it would be measured in a negative direction from the origin O ; that is to say, from right to left. Since this is equivalent to the operation of multiplying the component by -1 , we say $j^2 = -1$ and, therefore, $j = \sqrt{-1}$. The square root of -1 is an imaginary mathematical quantity and has no significance other than the operational one we have assigned to j . Because j has this imaginary value we call the axis to which it refers us, and which is at right angles to the axis of reals, the *imaginary axis*.

Let us now consider the complex expression for a capacitive or negative impedance Z_c . The component due to capacitive reactance is $1/2\pi fC = X_c$. Since X_c would make an angle of -90 degrees with the axis of reals it is expressed in complex notation as $-jX_c$, or $(-j/2\pi fC)$. The complex expression for an impedance consisting of a resistance and a capacity in series is, therefore, $Z_c = R - (j/2\pi fC)$, or simply

$$\bar{Z}_c = R - jX_c.$$

For a positive and negative impedance in series we would have

$$\bar{Z}_A = \bar{Z}_L + \bar{Z}_c$$

Since $Z_L = R_L + jX_L$ and $Z_C = R_C - jX_C$:

$$\begin{aligned} Z_A &= R_L + jX_L + R_C - jX_C \\ &= (R_L + R_C) + j(X_L - X_C) \\ &= R_A + jX_A \end{aligned}$$

Whether j will have a positive or negative sign in the final result (after the values have been inserted) will, of course, depend upon whether X_L or X_C is greater. If X_L is larger, $X_L - X_C$ will be positive, so that j will be positive. If X_C is larger, j will obviously be negative; that is, $\bar{Z}_A = R_A - jX_A$. When $X_L = X_C$, then $\pm jX_A = 0$ and we obtain $\bar{Z}_A = R_A$, the two reactive components, which are in opposite phase, being equal. In that case the current is in phase with the applied voltage and the circuit is, as we know, in resonance. The expression for the current,

$$I_{r.m.s.} = \frac{E_{r.m.s.}}{\sqrt{R_A^2 + X_A^2}}$$

is then simply

$$I = \frac{E}{R}$$

To illustrate the application of the method of complex quantities to problems dealing with parallel circuits, let us consider the problem shown in Fig. 66.

In our previous solution of this problem we depended upon the use of certain formulas, such as

$$G = \frac{R}{R^2 + X^2} \quad (88)$$

and

$$B = \frac{X}{R^2 + X^2} \quad (89)$$

which had already been derived and without which we could not have proceeded. In dealing with certain problems such formulas may first have to be derived. We shall see how they may be arrived at through the complex method of solution.

We are already familiar with the fact that impedance is a complex quantity, and that, when a number of impedances are connected in series, the resultant is the sum of their complex expressions and not the simple algebraic sum of their values. The expression for the resultant or equivalent impedance, where a number of impedances are connected in parallel is, as we have seen previously:

$$\frac{1}{\bar{Z}_A} = \frac{1}{\bar{Z}_1} + \frac{1}{\bar{Z}_2} + \frac{1}{\bar{Z}_3}, \text{ etc.}$$

This equation somewhat resembles that for a number of resistances in parallel, namely, $\frac{1}{R_A} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$, etc. with the important exception, however, that the impedances are not simple algebraic quantities. Substituting the complex expressions for impedances in the above equation, we get

$$\frac{1}{\bar{Z}_A} = \frac{1}{R_1 + jX_1} + \frac{1}{R_2 + jX_2} + \frac{1}{R_3 + jX_3}, \text{ etc.}$$

To get rid of j in the denominator we rationalize this expression by multiplying each complex fraction by its own denominator with its sign reversed, thus:

$$\begin{aligned} \frac{1}{R_1 + jX_1} \times \frac{R_1 - jX_1}{R_1 - jX_1} &= \frac{R_1 - jX_1}{R_1^2 + X_1^2} = \frac{R_1}{R_1^2 + X_1^2} - j \frac{X_1}{R_1^2 + X_1^2} \\ \frac{1}{R_2 + jX_2} \times \frac{R_2 - jX_2}{R_2 - jX_2} &= \frac{R_2 - jX_2}{R_2^2 + X_2^2} = \frac{R_2}{R_2^2 + X_2^2} - j \frac{X_2}{R_2^2 + X_2^2} \\ \frac{1}{R_3 + jX_3} \times \frac{R_3 - jX_3}{R_3 - jX_3} &= \frac{R_3 - jX_3}{R_3^2 + X_3^2} = \frac{R_3}{R_3^2 + X_3^2} - j \frac{X_3}{R_3^2 + X_3^2} \end{aligned}$$

Collecting these rationalized terms again we get

$$\begin{aligned} \frac{1}{\bar{Z}_A} &= \left(\frac{R_1}{R_1^2 + X_1^2} - j \frac{X_1}{R_1^2 + X_1^2} \right) + \left(\frac{R_2}{R_2^2 + X_2^2} - j \frac{X_2}{R_2^2 + X_2^2} \right) + \\ &\quad \left(\frac{R_3}{R_3^2 + X_3^2} - j \frac{X_3}{R_3^2 + X_3^2} \right). \end{aligned}$$

We see that the real components which are of the form $\frac{R}{R^2 + X^2}$ are what we previously called the conductance g and that the imaginaries, which are of the form $\frac{X}{R^2 + X^2}$, represent the susceptance b . Collecting the reals and imaginaries, we may, therefore, write the above expression as

$$\frac{1}{\bar{Z}_A} = (g_1 + g_2 + g_3) - j(b_1 + b_2 + b_3) = G_A - jB_A$$

where G_A = the resultant or sum of the real components, B_A = the sum of the imaginaries.

Since the reciprocal of the impedance $1/\bar{Z}$ is called the admittance \bar{Y} , we have, $\bar{Y}_A = G_A - jB_A$; and, since the magnitude or value of a complex expression is given by the square root of the sum of the squares of its real and imaginary parts (components), we get $|\bar{Y}_A| = \sqrt{G_A^2 + B_A^2}$. This was another of the formulas previously used in solving the problem

in connection with Fig. 66. The numerical solution of our problem has already been given (see Sec. 66).

We see that $G = \frac{R}{R^2 + X^2}$ can have only positive values regardless of whether susceptance $B = \frac{X}{R^2 + X^2}$ is positive or negative, depending on the sign of X . Parallel susceptances must therefore be added algebraically to take their sign into account. Admittances cannot be added directly, however, since they are complex. It is to be noted that the j term in the expression for admittance, $\bar{y} = g - jb$, or $\bar{Y}_A = G_A - jB_A$, is negative only when the circuit is predominantly inductive, that is, when the susceptance is positive. When the susceptance, or sum of the susceptances are negative, that is, the circuit is predominantly capacitive, the complex expression becomes $\bar{y} = g + jb$ or $\bar{Y}_A = G_A + jB_A$. The sign in the case of susceptance is therefore directly opposite to that in the case of impedance $\bar{Z} = R + jX$, which is positive when inductive; and negative, $\bar{Z} = R - jX$, when capacitive.

69. Power Relations in Alternating-current Circuits.—It has been shown that the true power in any circuit in which the current and voltage are in phase is the product of $E \times I$ or $I^2 \times R$. That this power is available for heating, lighting, and operating machinery is apparent. However, in alternating-current reactive circuits the true power cannot possibly be the product of $E \times I$, since the current and voltage are not in phase with each other. That is to say, the reactive parts of the circuit, such as the inductance or the capacity, actually return energy to the supply owing to conversion from potential to kinetic energy. Consequently, the true power in an alternating-current circuit must be dependent upon the ratio of the resistance component to the impedance component of the circuit as well as the product of EI , or the apparent power component. Hence, for single-phase circuits:

$$\text{True power} = \text{apparent power} \times \frac{\text{resistance}}{\text{impedance}}$$

or

$$\text{Power} = E \times I \times \frac{R}{Z} \quad \text{watts} \quad (92)$$

Furthermore, since the ratio R/Z in a right triangle is called the cosine of an angle, or $\cos \theta$, and also the *power factor*, the power formula may be similarly expressed

$$\text{Power} = E \times I \times \cos \theta \quad \text{watts}$$

Figure 68 illustrates the power curve in an alternating-current circuit containing inductance.

It will be noted that the power curve falls to zero at the points at which the current or voltage is zero. That is to say, at any instant when either the voltage or current is at zero the instantaneous power will also be zero. Obviously, therefore, since the current and voltage do not reach the zero points of the cycle in a reactive circuit simultaneously or in phase, the power must lie at some point between their respective zero points. Hence, the effective power $E I \cos \theta$ must depend upon the energy expended in the circuit during the cycle.

It will be noted that the frequency of the power curve is twice that of the current; also, that the power portion of the curve below the horizontal

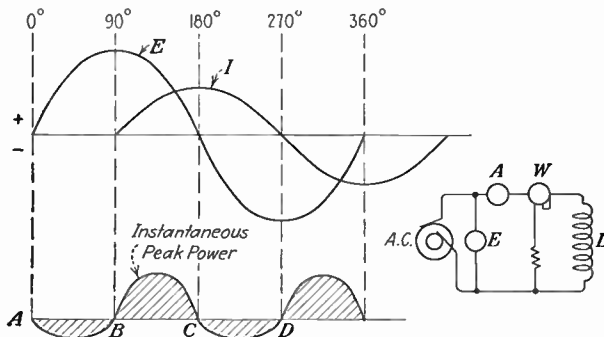


FIG. 68.—Voltage, current, and power relations in an alternating-current circuit.

axis at the points AB and CD represents the amount of energy returned to the circuit by the kinetic energy of the collapsing magnetic field.

The condition of power expenditure as just explained applies to common alternating-current circuits of a sinusoidal character and does not apply to waves of a non-sinusoidal character.

70. Kilovolt-amperes.—The rating of alternating-current machinery, such as transformers and alternators, depends upon the amount of heat the windings can stand. This in turn is dependent upon the flux which must be produced in the machine. Flux in turn is dependent upon the voltage and current in the windings and is independent of phase relations. It is logical, therefore, that the power factor should be neglected in computing the load an alternating-current device can carry, and this is done. These machines are rated in kilovolt-amperes which is a direct measure of the heating factors in the windings and a true measure of the capacity of the machine to do work. The abbreviation for kilovolt-amperes is kva.

Thus,

$$\text{Power rating} = \frac{E \times I}{1,000} \quad \text{kilovolt-amperes} \quad (93)$$

71. Polyphase Alternating-current Circuits.—As the only polyphase circuits in common use today are of the three-phase system, our discussion will be centered on three-phase equipment.

If a revolving-field alternator has three equispaced armature windings, as shown in the elementary diagram (Fig. 69), the induced voltage will

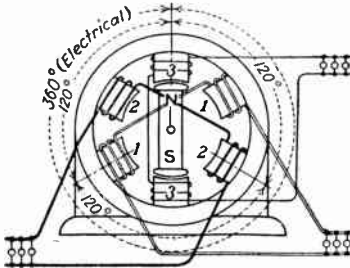


Fig. 69.—Revolving-field alternator.

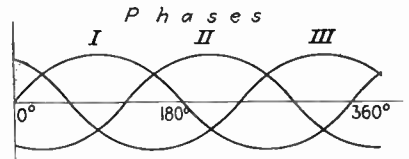


Fig. 70.—Voltage relations in a three-phase system.

be of a three-phase characteristic, each phase following the other by 120 degrees, as shown in Fig. 70. This machine, as connected to the load shown in the figure, is a simple balanced three-phase system. Such a generator would require six terminals, two for each phase, and is now obsolete. Other methods of connection, called the *delta* and *star* systems, are now employed which make only three terminals necessary on the apparatus. The star system of connection is also called the *Y* or *wye* method. These systems are illustrated in the diagrams of Fig. 71.

Current and voltage relations in a three-phase alternator depend upon the method of connection employed, that is, whether it is delta or *Y* connected.

Delta- or Mesh-connected Alternator.—*a.* The scheme of armature-coil connection for this circuit is shown in Fig. 71(a). The windings in the machine are not, of course, arranged in the position shown in the diagram, but the electrical circuit is schematically as shown. In an actual machine it is very difficult to tell the method of connection employed, as to do so requires penetration of the insulating wrapping around the windings which is sometimes a very difficult operation.

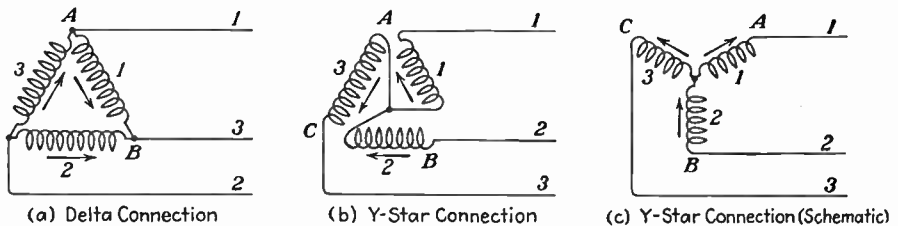


Fig. 71.—Methods of connecting electrical apparatus to a three-phase system.

b. In a delta-connected machine the voltage relations in coils 1, 2, and 3 may be shown vectorially as in Fig. 72. Note that no current can circulate in the armature because the voltage drop across each coil is exactly balanced by the combined voltage

drop of the other two coils. However, if three lines are taken off the armature and numbered 1, 2, and 3, as shown in Fig. 71(a), an e.m.f. will exist between certain points in the three-wire system as will be seen from the following consideration.

c. Consider lines 2 and 3; a current is flowing (assuming a closed outside circuit) from line 2 to line 3 through the coil 2 and also through coils 3 and 1 which is one phase of the circuit. Now consider lines 1 and 3. It is seen that the e.m.f. across these lines is that of coils 1 and 2 added together minus the drop across coil 3. The same reasoning applies to lines 1 and 2 where the e.m.f. is the sum of the e.m.f.s. across coils 2 and 3 minus that produced in coil 1.

d. It is to be noticed that this last mentioned phase is negative if the other two are considered positive. It is further to be noticed that, at the instant described, the maximum voltage is across lines 2 and 3 because the combined e.m.f.s. of all three coils are across these lines, while, if the other two phases are considered, the polarity of one coil of the three is such as to make it negative in respect to the other two causing it to subtract a certain value from the total e.m.f. across the line considered. In studying these current and voltage relations reference should be made to Fig. 70 which shows

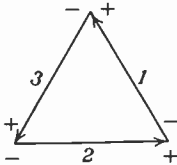


FIG. 72.—Voltage relations in the three coils of the delta connection.

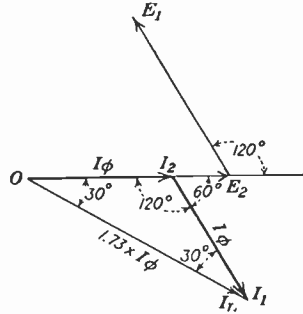


FIG. 73.—Vector representation of a three-phase circuit.

the three phases of e.m.f. and their relation to one another. In Fig. 72 the voltage relationship in the three coils of the delta connection are shown vectorially, and in Fig. 71(a) the delta connection is shown schematically with the current flow indicated by arrows.

e. In a delta-connected alternator the voltages across all phases are equal and have an angular relationship of 120 degrees in a balanced circuit as shown in the vector, Fig. 72. Also, the voltage across any two wires of a delta-connected three-phase system is equal to the voltage in one phase coil of the armature.

f. The current delivered by each phase to the line wire is equal, and all have identical phase relationships to their voltage component, that is, if the lag or lead in one phase is a certain amount, then the same lag or lead exists in the other phases. The power factor for a three-phase circuit can, therefore, be expressed by one figure as was previously explained.

g. Let us consider now the total current flowing in a line of a three-phase delta circuit. For example, take line 3 [Fig. 71(a)]; current in this line is supplied by coils 1 and 2 because these are the only two coils connected to this line. It will now be seen that if we draw a vector of the current relations between the two coils 1 and 2 which are supplying current to line 3, the current in coil 2 is assumed to flow in the same vectorial direction as the voltage. In Fig. 73 this voltage is shown as the light line which terminates in the arrowhead marked E_2 ; the current supplied by this coil is shown as the superimposed heavy line terminating at I_2 .

h. The voltage in coil 1 is shown by the light line extending upward from E_2 terminating at E_1 . Now here is an important point to notice; the current in coil 1, in order to supply current to line 3, must flow toward line 3, or in the same direction as the current in coil 2 which is in the opposite direction to the voltage in coil 1 (E_1). It must then be shown as opposite in the vector. Now the current in coil 1 is additive to the current in coil 2, and, therefore, the vector line representing this current must start where the current I_2 stops; it is, in effect, added to this current.

i. Now, how long shall this line be? To draw a vectorial line, three things must be known, namely, where does the line start and where does it stop, and in what direction does the line go? This boils down to, what is the length of the line and at what angle is it plotted?

j. As the currents supplied by each phase (coil) to the line are equal, the length of the vectorial line I_1 must equal the length of the vectorial line I_2 (see Fig. 73). As there is a 120-degree phase difference between I_1 and I_2 , the line I_1 must be plotted so that it forms a 120-degree angle with I_2 .

k. The current in line 3 or I_L is the vectorial sum of I_1 and I_2 . To the student familiar with trigonometry,

$$I_L = 2 \times I_2 \cos 30^\circ = \sqrt{3} \times I_2$$

or

$$I_L = \sqrt{3} \times I_\phi \quad \text{or} \quad 1.73 \times I_\phi$$

where I_ϕ is the current in each phase and I_L the current in each line. To the practical man not familiar with trigonometry the relationship may be expressed:

$$\text{Line current } I_L = 1.73 \times I_\phi \tag{94}$$

l. It will be seen by studying Fig. 73 that, while each phase current is in phase with the e.m.f. which produces it, the total line current I_L and the phase currents producing it are separated by a 30-degree phase angle.

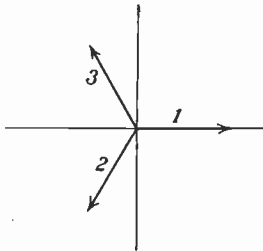


Fig. 74.—Voltage relations in a star connection.

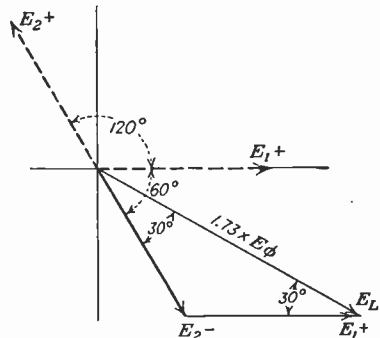


Fig. 75.—Vector representation of voltages in a star connection.

Y or Star Connection.—a. This method of connection is shown in Fig. 71(b) and (c), where three like ends of coils 1, 2, and 3 are connected together, and the free ends are connected to the three line wires. The voltages in the armature windings 1, 2, and 3 are shown in the vector diagram Fig. 74. As the phase angle between the voltage and current in each phase is the same, this same figure can also represent the current relationship in the three phases. For example, the voltage in coil 2 is always 120 degrees behind that in coil 1, and the voltage of phase 3 is 120 degrees behind phase 2. This is vectorially shown by the lines 1, 2, and 3 in Fig. 74. Also, in this

same figure, line 2 properly shows the 120-degree difference between phase 2 and phase 1, similarly between phase 3 and phase 2.

b. Referring back to the schematic diagram of a Y-connected alternator [Fig. 71(b) and (c)], it is seen that the current in line 1 must come from coil 1, that in line 2 from coil 2, and that in line 3 from coil 3. The word *coil*, as used here, and as shown in the diagrams, means an entire phase winding or *belt*.

c. The e.m.f. across any phase of the line is the algebraic sum of the e.m.f.s. produced by the two phase windings across which the line is connected. For example, in the Y-connected alternator [Fig. 71(b)], the e.m.f. across lines 1 and 2 must be produced by coils 1 and 2 because these are the only two coils in the circuit. Now, if the relative direction of the e.m.f. in each coil is as is indicated by the arrows in the figure, we may plot a vectorial diagram of the total voltage in each phase as shown in Fig. 75. Here it is seen that the e.m.f. produced by coil 1 is plotted as vector line E_1 and represents a positive voltage. Assuming that line 1 is positive, at the instant of time represented by the vector, then the line 2 must be negative. Now to get from line 2 (negative) to line 1 (positive), we must pass in a reverse direction through coil 2. Therefore, if, on the vector, the positive e.m.f. in coil 2 is represented by the dash line, then a reverse of this effect, or negative e.m.f., would have to be in the reverse direction to the dash line, or vector line E_{2-} . The total e.m.f. across lines 1 and 2 would be the vectorial sum of E_{1+} and E_{2-} , which gives us vector line E_L .

Expressed mathematically,

$$E_L = 2E_{1+} \times \cos 30^\circ = \sqrt{3} \times E_\phi = 1.73 \times E_\phi \quad (95)$$

where E_L is the voltage across the line E_L , and E_ϕ is the voltage in one phase.

Summary.—Voltage and current relations in a three-phase system may be summarized as follows:

a. In any three-phase system the voltage across any one phase has a phase-angle relationship of 120 degrees to the voltages in the other two phases.

b. In a balanced delta- or mesh-connected system the voltage across any line of the system equals the e.m.f. across any armature winding or belt in the alternator, $E_L = E_\phi$. The current in each line equals 1.73 times the current in each armature winding, $I_L = 1.73 \times I_\phi$.

c. In a balanced Y- or star-connected system the current across any line of the system equals the current in the armature path to which the line is connected, $I_L = I_\phi$. The voltage in each line equals 1.73 times the voltage summation of the armature windings to which the line is connected, $E_L = 1.73 \times E_\phi$.

72. Power in a Three-phase System.—In a simple three-phase balanced circuit, as shown in Fig. 69, the power in the circuit would be

$$P = 3 \times E_\phi \times I_\phi \times \cos \theta \quad \text{watts} \quad (96)$$

where $E_\phi = E_L$ which is the line voltage.

$I_\phi = I_L$ which is the line current.

$\cos \theta =$ the angle between E and I , or the power factor.

However, in a delta-connected machine,

$$I_{\phi} = \frac{I_L}{\sqrt{3}} \quad \text{and} \quad E_{\phi} = E_L$$

Therefore,

$$I_L = I_{\phi} \times 1.73$$

and in a Y- or star-connected machine

$$E_{\phi} = \frac{E_L}{1.73} \quad \text{and} \quad I_{\phi} = I_L$$

Therefore,

$$E_L = E_{\phi} \times 1.73$$

Consequently, power in a delta- or Y-connected machine is

$$P = 1.73 \times E \times I \times \cos \theta \quad \text{watts} \quad (97)$$

where E and I are line voltage and current, respectively.

Example: A three-phase transformer is connected across a three-phase line in which each phase draws 10 amp. The voltage across each phase is 220 volts. The power factor is 0.85. What are the true and apparent power in the circuit,

Solution: True power = $1.73 \times 220 \times 10 \times 0.85 = 3,235$ watts.

Apparent power = $1.73 \times 220 \times 10 = 3,806$ or 3.8 kva.

CHAPTER III

INTRODUCTION TO VACUUM TUBES

The triode or three-element vacuum tube as used in radio communication consists of a filament, a grid, and a plate, all sealed in an evacuated bulb, which is provided with the necessary external terminals for making contacts with the three elements inside. A study of the fundamentals of the triode will give an understanding of vacuum-tube operation on which a knowledge of the action of the tetrode, pentode, and similar multielement tubes can be built. For this reason, therefore, the triode will first be studied in considerable detail; a study of other types of tubes will then follow logically.

I. VACUUM-TUBE CHARACTERISTICS

73. The filament is simply a thin wire made up of a special composition which becomes heated when a current flows through it. Some filaments must be heated to incandescence (white hot) for the best results, while others perform even more satisfactorily when operating at a dull-red temperature. The two most important types of filaments used in the modern radio vacuum tube are thoriated tungsten and platinum coated with oxides of barium and other elements. The former usually operates best when the filament is white hot while the latter operates most efficiently when in a state of dull or cherry red. The filament may be looked upon as a substance which evaporates when heated and then gradually disappears. This may be very readily understood if we use water as an analogy for evaporation. For example, at ordinary temperatures water is known to evaporate very slowly, but, as it becomes heated, more and more evaporates until finally there is a violent boiling and the water is transformed into steam and passes into the atmosphere.

Similarly, when the filament in a vacuum tube is heated to operating temperature, the small particles of matter (electrons) which make up the physical structure of the filament are gradually escaping into space.

It has been shown that all substances are composed of infinitely minute particles of matter called *molecules*, and that these molecules in turn are made up of still smaller particles called *atoms*. Now, at ordinary temperatures, these small bodies are constantly colliding with one another, and consequently, if the temperature is raised, the motions of these molecules speed up and they start to bombard their neighbors with such tremendous velocity that the surface tension, due to the forces of cohesion between the atoms of a solid, is diminished, and some of the free electrons

in the materials may actually be forced beyond the surface of the material and ejected into space. It can readily be seen that this action may produce in the immediate vicinity of the filament a vapor, or cloud, similar to the evaporation of the water when boiling. This cloud is found to be in the form of a negative state of electricity due to the fact that the entire area near the filament is made up of the free electrons which have broken through the surface of the material. The reader must, therefore, always picture the immediate vicinity of the filament, when it is heated to operating temperature, as an atmosphere charged with electrons or negative particles of electricity.

This ever present negative charge (called *space charge*) or field in the immediate vicinity of the heated filament gives rise to one of the most important fundamentals in vacuum-tube theory. As stated before, if a

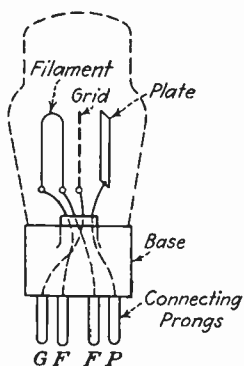


FIG. 76.—Three-element vacuum tube (triode).

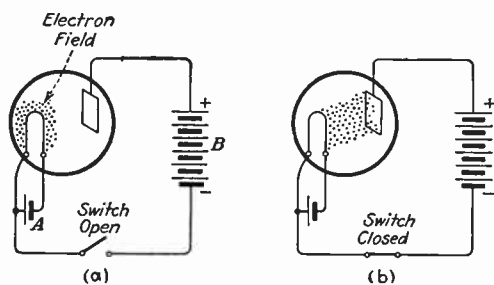


FIG. 77.—Diagrams to illustrate the effects of the space charge in a vacuum tube.

positively charged body is placed in the vicinity of a negatively charged body, an attraction will result. In other words, electrons will move in the direction of the positively charged body. Now, if we apply this principle to the vacuum tube, let us see what happens when a positive charge is placed on a metal plate located near the filament inside of the vacuum tube.

Figure 76 illustrates the construction of a simple vacuum tube. Figure 77(a) illustrates the effect obtained in the immediate vicinity of the filament, when no positive potential is applied to the plate (plate switch open). Note the cloud of electrons in the vicinity of the filament.

The function of battery A is merely to apply a voltage pressure across the filament terminals to force a current to flow through the filament to produce the necessary heating effect. This pressure may be supplied by a battery, generator, or a transformer, provided the proper value of voltage is applied. Too much voltage may cause an excessive amount of current to flow through the filament and consequently cause it to evapo-

rate too quickly or, in other words, to burn out. Every tube must operate at the proper voltage as specified by the manufacturer.

It will be seen from Fig. 77(a) that no electrons are attracted to the plate, because the plate potential is not applied and therefore the electrons concentrate around the vicinity of the filament. However, as soon as the plate-circuit switch is closed [Fig. 77(b)], a positive potential will be applied to the plate which causes the electrons to be drawn over (attracted) to the plate. This is called the *Edison effect* and gives rise to an electron flow in the plate circuit in the direction of the arrows as illustrated in Fig. 78.

74. Effect of Plate Potential on Plate Current.—The filament sends forth its electrons at a certain rate, depending upon the potential of the plate itself in respect to the filament and upon the degree of heat at which the filament is burning. It is easily conceivable that there will be a

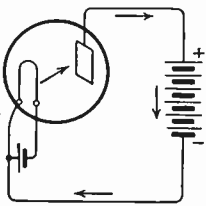


FIG. 78.—The arrows indicate the direction of electron (current) flow in a vacuum tube.

point, assuming that the plate has a high potential, when it will be attracting all of the electrons which leave the filament. When this happens, an increase in plate potential cannot result in a heavier electronic flow from the filament because all of the electrons leaving the filament are already reaching the plate. Hence, increasing the plate potential beyond this point will not result in a larger plate current. (Remember, the amount of current flowing in the plate circuit is governed absolutely by and really is the electronic stream from the filament to the plate.) This is known

as the *saturation point*, and the electronic stream cannot be increased beyond this point.

Now then, let us assume that the filament is burning at its rated brilliancy and sending forth a maximum number of electrons. Assume the plate potential to be adjusted low at this particular instant. It will be readily seen that all of the electrons leaving the filament will not be attracted to the plate: first of all, because the plate potential is not sufficiently high to draw them all over, and second, because the electrons have a tremendous repelling tendency. In other words, some of the electrons leaving the filament are constantly repelled back, thus leaving a certain number which really never get over to the plate. This latter condition always holds true even though the plate potential is increased. Of course, the greater the value of the plate potential, the more electrons will be attracted.

An increase in plate potential will increase the plate current up to a certain point beyond which any further increase of plate potential will have no effect upon the plate current. This is the saturation point and depends upon the design characteristics of the tube. It is always true that a reduction of the plate potential will decrease the plate current as

soon as the potential of the plate falls so low that the intensity of the electron stream from the filament to the plate is reduced.

In practice, it is desirable to maintain the filament temperature constant by adjusting the filament voltage to that recommended by the manufacturer and not to vary the plate current by varying the filament temperature. Plate current should be increased or decreased only by variation of the plate-supply potential, or the grid-bias voltage, as will be seen later.

75. Effect of Filament Current on Plate Current.—This effect was discussed in the preceding explanation, but in order that it may be clearly understood let us assume a practical example.

Suppose the plate is at a potential great enough so that it can attract 2,000 electrons per second from the filament. Suppose the filament is emitting 1,000 electrons per second. Each electron as it leaves the filament carries a negative charge, and, therefore, the space between the filament will be charged negatively to the value of 1,000 units, while the plate is charged positively to the value of 2,000 units. The result is that the plate charge is stronger than the filament space charge, and all of the electrons which leave the filament per second are attracted to, and reach, the plate.

Suppose, further, that the filament temperature is raised so that 1,500 electrons are emitted. The space charge is then 1,500 units negative and the plate, still remaining at 2,000 units positive, attracts nearly all of the electrons emitted by the filament.

Suppose now the filament temperature is raised so that 2,000 electrons are emitted making the space charge 2,000 units negative as against the plate charge of 2,000 units positive. The plate still attracts all of the 2,000 electrons emitted by the filament, but the saturation point for the plate potential has been reached; the space charge now equals the plate charge.

Up to this point there has been an increase in plate current for every increase in filament current because the path (the electron stream) between the filament and the plate has been increased. Now suppose that the filament temperature is increased so that 2,500 electrons are emitted. The space charge is 2,500 units negative as against the plate charge of 2,000 units positive. The result is that 500 electrons are repelled back to the filament; lost as far as their effect upon the plate current is concerned. Any further increase in filament temperature would be useless, as all the electrons over the 2,000 limit, which the plate will attract, are useless in effect. Therefore, increasing the filament current does not result in an increased plate current as long as the plate potential remains at the point where the plate attracts only 2,000 electrons.

If the plate voltage is increased, and the positive potential of the plate is raised so that more than 2,000 electrons are attracted to the plate, a

still further increase in plate current will result. Thus, a constant increase in plate current will result until all of the electrons are attracted to the plate, or, in other words, until the saturation point has been reached. In general practice, however, the filament current is never varied after the proper filament voltage has been applied. Hence, all changes in plate current must be made by a variation of the plate-supply voltage, and that only up to the point specified by the manufacturer. Tubes operated in excess of their designed plate- and filament-voltage ratings will operate inefficiently, and such operation may result in a considerable decrease of their normal operating life. A filament operating at a voltage of less than 10 per cent over its normal voltage rating will decrease to a value of 50 per cent of its normal operating period. The life of the various types of receiving and transmitting tubes is found to be between 1,500 and 5,000 hr., seldom more than 10,000 hr. It is good design practice, therefore, to provide a sufficient number of electrical indicating instruments (voltmeters, milliammeters, etc.) and fuses in vacuum-tube circuits, especially in transmitting circuits where operating conditions frequently change.

The preceding explanation should give the reader a fairly good idea of the relative effects of increasing the plate or filament potential, but he must not forget that the figures quoted were to make for simplicity of illustration only and that the electrons actually emitted may number in the millions. Furthermore, it should be born in mind that, even though the plate potential is at a point at which a maximum number of electrons are attracted (saturation), there are many electrons which never really reach the plate because of their tremendous repelling tendencies.

76. The Control Grid.—Bearing in mind that in the immediate vicinity of the filament there is a heavy negative field or space charge which is the true determinant of the amount of plate current that will flow when a positive potential is applied to the plate, let us see what would happen if a third element or grid is interposed between the filament and the plate.

The grid consists of very fine wires laterally arranged and spaced, so that a large portion of the electrons can pass through them. Now, just as the plate attracts electrons when it is positively charged, so will the grid attract electrons if a positive potential is applied to it.

Let us assume the grid to be connected to a variable potential, such as is illustrated in Fig. 79. It will be seen from this illustration that either a positive or negative potential may be applied by simply moving the sliding arm of the potentiometer (voltage-dividing resistance) either up or down from the center position. When the arm is in the central position, no e.m.f. is applied to the grid and, consequently, no effect is produced on the steady plate current as indicated by the plate milliammeter.

Now let us assume that the slider arm is moved toward the extreme positive end of the resistance. A certain amount of positive potential

will be applied to the grid depending upon the number of cells connected in series from the center tap of the batteries to the extreme positive (+) end. Assuming this to be three dry cells in series and each cell to have a voltage of 1.5 volts, the total positive voltage applied to the grid will be 4.5 volts. This voltage may be gradually decreased to zero by moving the sliding arm toward the center or zero point of the potentiometer.

Now, since it has been established that the positive potential of the plate will attract electrons, it is equally accurate that a grid interposed between the filament and the plate will also attract electrons when a positive potential is applied to it. However, owing to the fact that the grid is nearer to the filament than the plate, it will have a much greater effect on the electron flow than an equal charge on the plate. In other words, a positive grid in close proximity to the filament will cause a greater amount of electrons to flow toward the plate because of its tendency to "break up" or neutralize the heavy negative space charge present near the filament. If, on the other hand, the grid is made negative by sliding

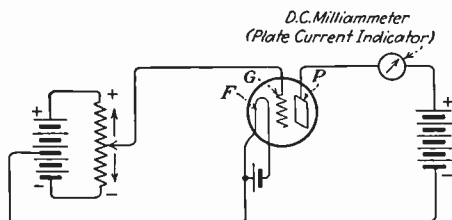


FIG. 79.—Circuit for providing variable bias to grid of vacuum tube.

the movable arm toward the negative side of the resistance, the space charge will not only appear again but, in addition, will tend to repel electrons which otherwise would reach the plate, consequently reducing the plate-current flow below its normal value. Normal value is the amount of plate current flowing with zero potential on the grid.

The amount of increase and decrease of plate current in milliamperes depends upon the value of the positive or negative charges applied. In any event, however, no matter how feeble the amplitude of the positive or negative grid e.m.fs. applied to the grid, a relatively larger effect on the plate current will result. Consequently even very feeble signal impulses will produce relatively large plate-current variations. In this action lies the ability of the three-element vacuum tube to amplify feeble signal-voltage variations. Incoming radio signals are of an alternating character, and, consequently, they apply to the grid voltages of an alternating character.

77. Vacuum-tube Characteristics.—In order that the operation of a vacuum tube might be clearly understood when operating into load circuits, such as the primary windings of radio-frequency or audio-frequency transformers, the tube should be visualized as an alternating-

current generator as illustrated in Fig. 80. Under these conditions a maximum power will be absorbed by the load resistance or impedance when the load is equal to the generator resistance or impedance. That is to say, if the load resistance R_o or impedance Z_o is matched to the generator impedance, the power transfer from one circuit into another circuit, such as from the primary winding to the secondary winding of a transformer, will be a maximum if these conditions are fulfilled.

In vacuum-tube practice, however, these conditions are not always desired, particularly in alternating-current audio-frequency amplifying circuits, owing to the fact that an exact matching of the impedance to the plate resistance may produce serious distortion of the audio frequencies. For this reason certain audio-frequency load circuits may have an impedance much greater or smaller than the tube (generator) impedance, depending upon the type of tube used. This is one of the most important fundamentals in engineering and it must be thoroughly understood to obtain a clear picture of vacuum-tube operation.

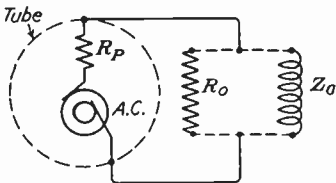


FIG. 80.—The vacuum tube and its load are analogous to an alternator and its load.

The general characteristics of vacuum tubes are most commonly referred to as the amplification factor of a tube, or μ , the alternating-current plate resistance, R_p , (equal to the alternating-current generator resistance in the analogy), and the mutual conductance, g . These are the most important characteristics which, in conjunction with the load conditions, determine the operating efficiency of the tubes.

78. Amplification Factor.—The amplification factor μ depends primarily upon the distance between the plate and the grid and also upon the number of, and spacing between, the laterals constituting the grid, and the size of the grid wire. The greater the distance between the grid and the plate, the greater will be the amplification factor of a tube. This factor remains practically constant with a variation of the plate potential and may only decrease slightly at low plate voltages.

The amplification factor may be defined as the ratio of a change in the plate voltage (necessary to give a change in plate current of a certain value) to a change of grid voltage (which will produce the same change in the plate current starting from the same initial conditions). Although the μ decreases somewhat at lower voltages, it is almost entirely dependent upon the geometry of the tube, and is therefore a constant, called the *voltage amplification factor* or *constant* of the tube.

It is very easy to understand how this factor may be determined if the elementary function of the grid as a space-charge control is thoroughly understood. For example, when the filament is heated to the proper emitting temperature, the free electrons which are emitted find themselves

under the influence of the electrostatic field created in the vicinity of the filament by the positive charge on the plate and the positive or negative charges on the grid. It will be remembered that the electron movement to the plate is a direct result of the charged fields set up by the plate and grid charges, respectively. Hence it may be seen that, through a variety of possible combinations of the plate and grid potentials, the number of electrons moving to the plate will be the same, regardless of what particular combination is used.

For example, if the plate potential is decreased causing a decrease in plate current, it will again be possible to restore the plate current to normal by making the grid potential more positive or less negative. Similarly, if the plate voltage E_p (symbol representing plate voltage) is increased or decreased by a small amount of dE_p (where dE_p represents a certain plate voltage change) causing a variation dI_p (where dI_p represents a change in plate current), the same plate-current variation may be obtained by keeping the voltage constant and varying, in the same direction, the grid voltage E_c (symbol representing grid voltage) by an amount dE_c , which will produce the same plate-current variation as did the plate-potential change dE_p .

This ratio is expressed by the formula

$$\mu = \frac{dE_p}{dE_c} \quad (98)$$

Hence, to determine the amplification factor of a tube, apply a certain voltage to the plate of the tube, say 45 volts, and note the amount of plate current flowing at that potential. The grid potential during this operation must be zero. Now decrease the plate potential to about 35 volts and note the plate-current reading. Let us assume the plate current at 45 volts to be 2 milliamperes (ma.) and 1 ma. at 35 volts. This constitutes a decrease in plate current of 1 ma.; which is a difference of 1 ma. Hence, the amplification factor (μ) is equal to the change in plate voltage divided by the change in grid voltage. Thus, if a negative grid voltage produced the same decrease of 1 ma. as did the change of plate voltage from 45 to 35 volts, then, by noting the value of the grid voltage which produces this change, the amplification factor can readily be determined by the formula for μ

Assuming the grid voltage which produced this change to be 2 volts negative, then, by application of the above formula the amplification factor will be found to be 5.

$$\frac{dE_p}{dE_c} = \frac{10}{2} \quad \text{or} \quad 5$$

Hence,

$$\mu = 5.0$$

79. Plate Resistance.—The plate resistance of a tube as usually specified in tube charts does not refer to the ordinary direct-current resistance between the plate and the filament but is the resistance offered to the flow of an alternating current (the signal variation) in the plate circuit. The plate resistance of a tube may be readily calculated from the values previously found for the amplification factor. The formula for finding the plate resistance is as follows:

$$R_p = \frac{dE_p}{dI_p} \quad (99)$$

where dE_p = change in the plate voltage.

dI_p = change in the plate current.

Hence, in the previous example the change dE_p in the plate voltage was 10 volts, and the change dI_p in plate current 1 ma., or 0.001 amp. Thus, from the above formula the ratio of 10:0.001 will be found to be 10,000, which is the alternating-current plate resistance in ohms.

80. Mutual Conductance.—One of the simplest methods for determining the amplifying efficiency of a vacuum tube is to compare the ratio of the amplification factor to the plate resistance of the tube. This is known as the *mutual conductance* of the tube and is found by determining the effects of the interrelation between grid voltage, plate voltage, and plate current, which may be computed by the ratio

$$G_m = \frac{dI_p}{dE_g} \quad \text{or} \quad \frac{\mu}{R_p} \quad (100)$$

In other words, the mutual conductance of a tube may be defined as the ratio of a small change in the plate current to the change in grid voltage necessary to produce the same change in the plate current.

81. The Alternating-current-operated Vacuum Tube.—The current flow in the plate circuit of a vacuum tube must be as nearly unvarying in amplitude as possible in order that no varying magnetic field will be produced about the output-load unit, such as the primary winding of an audio-frequency transformer. If the plate- and filament-supply voltages are obtained from batteries, the plate-current flow through the primary winding of the transformer represents a steady unmoving magnetic field. Consequently, if the magnetic field about any coil is stationary, no e.m.f. can be induced into the secondary circuit and to the grid of the succeeding amplifier tube. If, on the other hand, the plate voltage is supplied by a rectifier system in which the alternating current is changed to direct current through the medium of rectifier tubes, a very efficient filtering system must be provided to smooth out the direct-current pulsations, or ripples. This may be effectively accomplished by using a filtering system as described in the sections on ripple filters (Chap. XIII),

a. *The filament-type tube*, when operated from an alternating-current supply, requires that necessary provisions be made to maintain a steady potential between the filament and the plate. That is to say, the filament must be heated with an unvarying potential, so that the electron stream to the plate will be unaffected and steady. This condition might be obtained by heating the filament with a rectified and filtered alternating current. However, in view of the fact that the rectifier unit and filter coils would have to carry the filament currents it would necessitate the design of large current-carrying rectifier tubes and filter chokes. This procedure would not only make the cost prohibitive but impractical.

If, therefore, the filament is to be operated directly from an alternating-current supply source such as a transformer, means must be provided to minimize the effect of the alternating-current field upon the steady plate-current flow and the unsteadiness of the filament potential with respect to the plate.

For example, Fig. 81 illustrates a simple vacuum-tube plate circuit in which the plate voltage is assumed to be a well-filtered supply from a rectifier system which maintains a steady voltage upon the plate of the vacuum tube. The filament is lighted by an alternating potential from the transformer.

Now, since the negative lead of the plate-supply system must be connected to the filament to complete the path of the plate-current flow, let us see what would happen if this lead is connected to the point *A* or *B* of the filament circuit. A plate-current flow would result in the form of an electron stream over to the plate, through the voltage-divider resistance and back to the filament, completing the circuit. However, since the filament potential at the points *A* and *B* is constantly reversing in polarity, due to the alternating potential of the transformer, the polarity of point *A*, to which the negative plate-supply lead is connected, is constantly changing. This produces a slight variation of the electron stream flowing to the plate of the tube at the frequency of the alternating-current reversals at the tube-filament terminals. Now, since this unbalanced condition produces a slight effect upon the plate-current flow, a moving magnetic field will be present around the transformer primary winding (shown here as a coil in the plate circuit), resulting in the inducing of an e.m.f. to the grid of the succeeding amplifier tube connected to the secondary of the transformer (not shown). The effect of this variation will then be amplified through the succeeding tubes and will finally be brought out through the audible reproducing system in the form of a hum.

Now, if the negative plate-supply lead is connected to some point of the filament circuit in which the potential is at all times zero, the electron stream to the plate would be unaltered, and a steady, or unvarying, magnetic field will result. Theoretically this condition could be accu-

rately obtained if the negative plate-supply return lead were connected to the exact center of the filament at the point X (Fig. 81). This condition, however, remains true only if the paths from each side of the point X are equal in length. Otherwise the electrical balance would not prevail. Since it is not practical to bring the physical connection of the

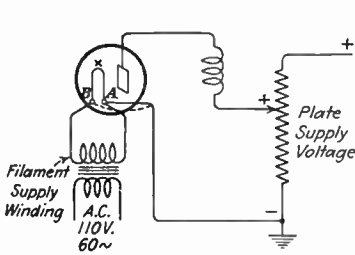


FIG. 81.—Illustrating effect of filament-return connection.

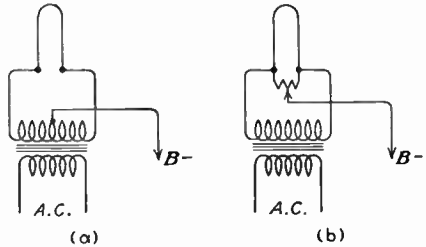


FIG. 82.—Two methods of connecting the filament-return to the filament.

B— lead to the exact center of the filament, two satisfactory arrangements which will produce the same effects are illustrated in Fig. 82.

It is of great importance to have the B— lead connected to the exact electrical center of the filament to obtain perfect balance, and for this reason the resistance method illustrated in Fig. 82 (b) is preferable. This method has a particular advantage in that the center tap may be varied to

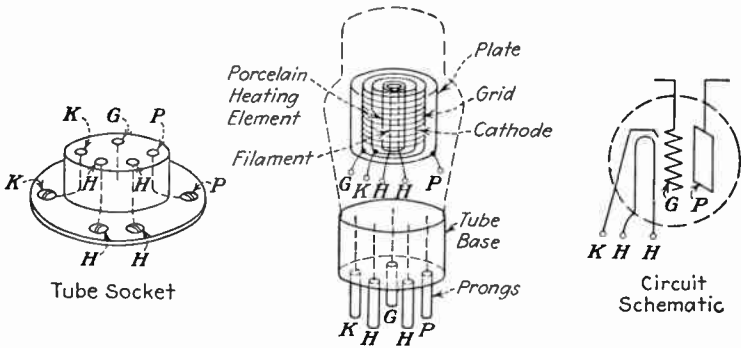


FIG. 83.—The heater-type vacuum tube.

compensate for any unbalance due to a slight difference in the length of the connecting leads.

b. The cathode-heater tube is shown in Fig. 83, which illustrates the internal structure of the cathode-heater or equipotential tube. The filament or heating element is designated by *HH*, the cathode or emitter by *K* or *C*, and the grid and plate by the usual *G* and *P*, respectively. The heater unit or filament is, therefore, not used as an electron emitter, but serves only to supply the proper heat to the emitting element, the

cathode. All of the elements are arranged in cylinders so that all points of the grid and plate are equidistant from the cathode thereby allowing an equal flow of electrons in all directions.

If an alternating e.m.f. of the proper potential is applied to the heater terminals of the tube, the porcelain heating element will gradually become heated. This causes the cathode cylinder of nickel oxide to become heated sufficiently to emit electrons. In other words, the cathode cylinder and not the heater is the emitter of the electrons to the plate. Hence, if the negative lead of the plate-power supply system is connected to the cathode instead of the filament, the potential difference between the *K* and *P* elements of the tube will be constant, even though the filament potential is varying at the alternating-current frequency.

The cathode-heater tube, therefore, will insure much less variation in the plate-current flow with alternating-current-operated filaments, thereby greatly reducing the possibility of an alternating-current hum.

There are, of course, other important factors which may contribute toward the elimination of an alternating-current hum in transmitters and receivers. For example, the filament wires carrying the alternating-current filament power must be very tightly twisted so that the magnetic fields between the wires will be canceled. Any varying magnetic fields situated near the grid or the plate wires may induce a sufficient alternating e.m.f. to produce an alternating-current hum.

When using the cathode-heater tube, the filament should also be center-tapped and grounded to eliminate stray capacity between the cathode and filaments. This is particularly important in transmitting circuits in which a slight potential difference existing between the filament and cathode may produce serious variations and modulations of the plate current. Perfect stability in all amplifying systems can only be obtained by the proper balancing and grounding of all filament and plate return circuits.

II. THE RADIO-FREQUENCY AMPLIFIER

82. General Amplifier Considerations.—The principles underlying the operation of audio- and radio-frequency amplifiers are the same, the only difference being that the frequencies to be amplified in the radio-frequency amplifier are considerably higher. Figure 84 illustrates a typical two-stage radio-frequency amplifier showing the grid and plate filtering system. Here the input voltages applied to the grid-filament circuit of the first radio-frequency amplifier tube are obtained from the parallel-resonant circuit L_2C_1 .

When an incoming signal e.m.f. is induced in the secondary circuit L_2C_1 at resonance, a voltage is built up across this parallel combination. The value of this e.m.f. is dependent upon one or more of the following conditions:

1. The step-up ratio of the radio-frequency transformer.
2. The impedance of the parallel combination.
3. The input impedance of the tube between the grid-filament capacity.
4. The resistance of the entire input circuit.
5. The reflected resistance or the mutual coupling M between the windings.
6. The grid biasing voltage.
7. The tuning of L_2C_1 .

It may be seen, therefore, that, if the ratio of turns in the radio-frequency transformer is larger in the secondary than in the primary, a greater e.m.f. will be applied across the secondary circuit at the points E .

If the impedance of the circuit L_2C_1 is kept large by maintaining a large L to C ratio, the e.m.f. across the parallel combination will also be considerably greater. That is to say, if the value of the capacity in the tuning condenser C_1 is kept as low as possible, and the value of L large, less current will circulate in the parallel-resonant circuit. Hence, the

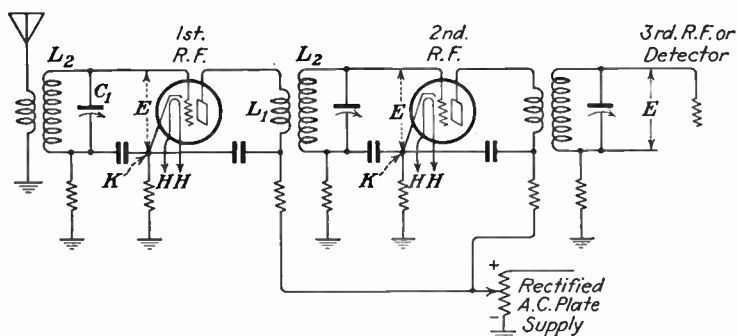


FIG. 84.—A typical two-stage radio-frequency amplifier showing grid and plate filtering system.

lower the current flow, the higher will be the effective impedance of the parallel circuit and the greater the voltage applied to the grid-filament circuit of the tube. This is an important consideration in the radio-frequency amplifier, particularly in receiving circuits, since the tube is to function primarily as a voltage amplifier. Thus, when very feeble signal voltages are being received, every available bit of e.m.f. must be secured to actuate properly the tube as a voltage-amplifying device.

The input impedance in a voltage-amplifying device must also be maintained as high as possible, particularly when small input voltages are being received. This is of great importance in the radio-frequency receiving amplifier because of the low signal voltages received.

The input impedance may be kept high by using a tube having a low grid-to-filament capacity and by preventing a flow of grid current during the period in which signal voltages are being received. This latter condition may be fulfilled by maintaining the proper value of the

negative C bias on the grid of the tube so that, when a positive potential is applied to the grid from the incoming signal, a minimum number of electrons will be attracted to the grid, thereby reducing the grid-current flow in the input circuit. However, since some current may flow in the input circuit, the direct-current and high-frequency resistance must be kept as low as possible to minimize the IR drops along the entire circuit.

Now, if the maximum voltage is built up across the grid-filament circuit of the first radio-frequency tube, the voltage amplification, A , of the tube may be computed from the general expression

$$A = E_2/E_1 \tag{101}$$

In other words, the higher the μ of the tube, and the larger the voltage applied, the greater the amount of amplification or gain that may be expected in the plate or output circuit of the tube. This is, however, a rather general condition since certain other factors must be included to obtain a more accurate result.

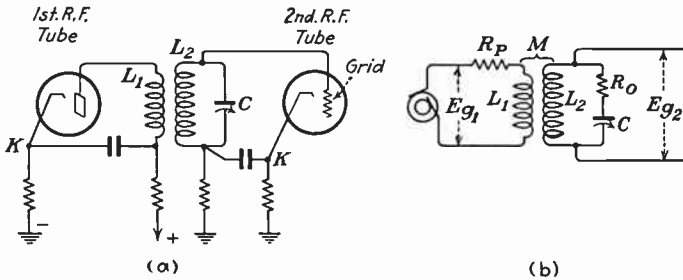


FIG. 85.—A simple radio-frequency amplifier circuit (a) and its equivalent (b).

Let us take, for example, the simple radio-frequency circuit in Fig. 85 (a) and reduce it to an equivalent electrical circuit [Fig. 85 (b)] containing a generator E_{g1} ($\mu \times e_g$), a primary and secondary inductance L_1 and L_2 , respectively, with mutual inductance due to the magnetic coupling M , and the primary- and secondary-circuit resistances R_p and R_o , respectively. A variable condenser C is shunted across the secondary to tune the circuit to a small band of radio frequencies.

The total voltage amplification or gain, A , of the tube and the transformer as a unit, therefore, must take into consideration all of the above factors. Hence, if it is assumed that the secondary circuit L_2C is adjusted to resonance, the total gain may be approximately computed from the ratio

$$E_{g2} = \frac{\mu E_g Z_o}{R_p + Z_o} \times \frac{L_2}{M} \tag{102}$$

Thus it may be seen that, for each frequency to which L_2C is adjusted, there is some value of mutual induction M which will give the

best results. This of course depends upon the conditions required, namely, selectivity or fidelity.

If it is desired to obtain an extremely selective response so that the voltage amplification at some one frequency is sharply defined as illustrated in Fig. 86(a), then the value of coupling chosen should approach zero. That is to say M must be very small. If, on the other hand, fidelity is the important consideration, such as in broadcast receivers where the frequency response curve must be wide enough to accommodate the modulated frequencies, the value of mutual induction M must be so adjusted that the response curve approximates Fig. 86(b).

Hence, by decreasing the mutual induction M between L_1 and L_2 , greater selectivity (sharpness of tuning) may be obtained, and, conversely, by increasing the mutual induction M , a broader response may be obtained. Thus, it may readily be seen that receivers which require fidelity and sensitivity at the same time must have their coupling adjusted to obtain a mean mutual (*mutual induction* is abbreviated to *mutual*) which will satisfy both conditions. In other words, if a fixed mutual

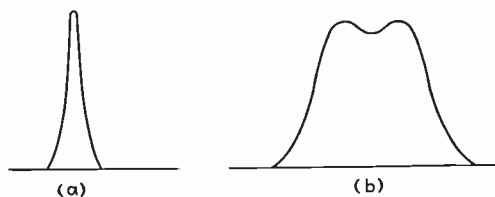


Fig. 86.—Frequency response curves.

is used, as in the case of most radio receivers, the value of M must not be too small because, when the value of the tuning capacity C is decreased so that the circuit L_2C is tuned to a high frequency, the selectivity becomes so great that serious distortion will result. If the condenser capacity C is increased to decrease the resonant frequency of the circuit, the tuning will be slightly broader, but the sensitivity will also be much less. Thus, when the tuning becomes much sharper at the higher frequencies, the fidelity is seriously jeopardized and a compromise must be effected between the higher and lower frequencies to obtain the most desirable results. In other words, the amplification and fidelity should be uniform over the entire frequency band covered by L_2C (see Band-pass Filters, Sec. 152).

If, on the other hand, the mutual is too high (tight coupling), the resistance introduced into the secondary may be too high, which will result in a decrease in selectivity and possibly a decrease in amplification. This resistance may be computed from the ratio

$$R_o = \frac{M^2 \times (2\pi f)^2}{R_p} \quad (103)$$

All amplifiers operating on the straight portion of the curve are commonly known as *class A amplifiers*. Their efficiency is low, as is their output, while their ratio of power amplification is high. (See Fig. 88.)

A simple method of calculating the gain of a radio-frequency amplifier would be to determine the load impedance Z_o into which the tube is working, the amplification factor μ of the tube, and its plate resistance R_p . For example, let us assume a direct-coupled tuned load across the tube resistance R_p , as illustrated in Fig. 87, using a triode amplifier of the -27 type. The circuit LC is adjusted to the frequency to be amplified, say 1,500,000 cycles. Now if the value of LC , together with the resistance R in the coil, has an impedance Z_o of say 20,000 ohms at resonance, $Z = L/CR$, then the amplification or voltage gain may be computed simply by the product

$$E_o = \mu \times E_g \times \frac{Z_o}{R_p + Z_o} \tag{104}$$

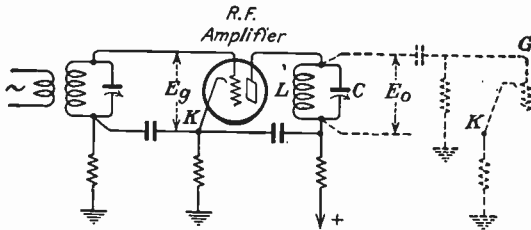


FIG. 87.—A direct-coupled tuned load across the output of a vacuum tube.

Since the plate resistance of a -27 type of tube is approximately 9,000 ohms at a plate potential of 135 volts and a grid bias of -9 volts, and the load impedance Z_o is generally adjusted to twice the tube impedance, the plate amplification will be approximately 6 times.

The expression 6 is simply a numerical value to show relative gain in amplifiers. If it is desired to compute the actual e.m.f., E_o developed across the load impedance Z_o , we may do so by the following reasoning: If the input signal e.m.f. at E_g is 0.001 volt, then the resultant plate-current variation I_p will be

$$I_p = \frac{\mu \times E_g}{R_p + Z_o}$$

Thus, the voltage developed across LC at E_o will be

$$E_o = \mu \times E_g \times \frac{Z_o}{R_p + Z_o} = 9 \times 0.001 \times \frac{18,000}{9,000 + 18,000} = 0.006 \text{ volt}$$

or 6 millivolts (approx.)

The gain of the circuit, therefore, is simply the ratio of the output e.m.f. E_o to the input e.m.f. E_g thus,

$$\text{Gain} = A = \frac{E_o}{E_g} = \frac{0.006}{0.001} = 6$$

The voltage and power gain in radio-frequency and audio-frequency circuits is generally calculated in decibels, db, as explained in Sec. 192.

83. The Class A Amplifier.—In the general treatise of radio- and audio-frequency amplifiers just completed, all the tubes were operated at the straight portion of the I_p - E_c characteristic curve, so that the output-wave shape in the plate circuit is a linear reproduction of the signal-wave shape. This condition is clearly illustrated by the curve in Fig. 88.

Although the efficiency of a class A amplifier can not exceed 50 per cent, as can readily be seen from the general formula for computing amplifier efficiency, the fact that this type of amplifier will give a symmetrical reproduction of the input signal voltage makes it highly desirable.

$$\text{Amplifier efficiency} = \frac{100 (E_p - E_{min}) I_p}{2 E_p I_{p(av)}} \quad (104A)$$

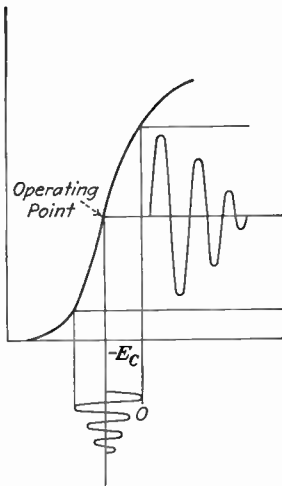


FIG. 88.—Class A amplifier operational curve.

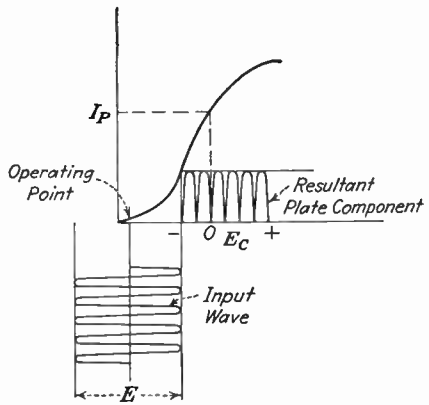


FIG. 89.—Class B amplifier operational curve. (Single tube)

where, $I_{p(av)}$ = average value of plate current.

I_p = peak value of plate current.

E_p = normal value of plate voltage.

E_{min} = minimum voltage applied to the plate due to the drop across the plate load impedance.

Proper operation requires that the tube be properly biased so that an incoming signal swing will not draw grid current during the positive halves of the cycle.

The class A amplifier should, therefore, always be used where the object is primarily voltage amplification and where power amplification

is of secondary consideration. Radio-frequency and audio-frequency amplifier systems for receiving purposes generally use the class A method in all stages preceding the final power-amplifier stage. However, many high-quality power-amplifier systems, particularly those used as speech amplifiers in broadcasting studios, use this system of operation in the final power-amplifier stage.

84. The Class B Amplifier.—This amplifier operates at the lower portion of the I_p-E_c characteristic as illustrated in Fig. 89. When a signal e.m.f. is applied across the grid-filament circuit of a tube, the plate variations will resemble a half-wave output owing to the adjustment of the grid bias to, or very nearly to, the point of plate-current cut-off. Hence, when an alternating e.m.f. is applied to the grid, the plate variations will take place only during the intervals in which the peaks of the signal or input voltage are positive. In other words, since no current is flowing in the plate circuit during the period in which no input voltage is applied, owing to the bias adjustment being at the blocking or cut-off point, no plate current can flow when a signal is being received during the negative halves of the cycle. This, of course, indicates quite definitely that the signal-wave shape is not faithfully reproduced in the plate circuit of the vacuum tube, owing to the loss of the negative half of the incoming cycle. In a radio-frequency amplifier, however, the full cycle may be reproduced in the plate circuit if an oscillatory or "tank" circuit is provided as illustrated in Fig. 90.

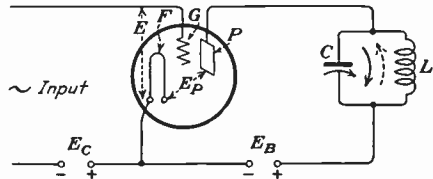


FIG. 90.—Tank circuit (LC) connected in plate circuit.

During the period in which the positive peaks of the incoming-signal alternating e.m.f. are applied to the grid of the tube, the potential difference across PF also applies an e.m.f. across the tank circuit CL since it is effectively in parallel to the tube generator PF . This charges the condenser C , which in turn discharges through the inductance L in the direction of the heavy arrow. That is to say, as soon as the positive peak in the plate circuit begins to diminish, the condenser discharges through L in the direction of the heavy arrow and discharges the condenser on the opposite plate. At this instant the positive peak in the plate circuit has disappeared owing to the reversal of the input grid e.m.f. to negative. During this negative period no current flows in the plate circuit of the tube, and no e.m.f. is built up across PF . However, in view of the charge that has been built up on the opposite plates of the tank condenser, the circuit will continue to oscillate in the direction of the dotted arrow thereby completing the other half of the wave lost by the negative signal cycle. This action is sometimes referred to as the

inertia or *flywheel effect* of an inductance because the inductance and capacity resemble a mass in motion against the action of a spring.

Hence, the class B system may be successfully used in radio-frequency amplifiers by using a single tube without jeopardizing the wave shape in the plate circuit owing to the flywheel action of the tank circuit.

In audio-frequency amplifiers, however, the class B system cannot be used with a single tube because no oscillatory or tank circuit is provided. Hence two tubes must be operated alternately to accommodate the complete cycle. This is frequently known as a *push-push* amplifier and differs from the so-called *push-pull* amplifier in characteristics only.

Figure 91 illustrates a push-pull or push-push audio-frequency amplifier. If this amplifier is so adjusted that it operates on the straight portion of the I_p-E_c characteristic curve it is known as a class A or push-pull amplifier. If the amplifier is adjusted so that it operates at or near the cut-off point, it is known as a class B or push-push amplifier.

When a tube is operating as a class B amplifier the output power is

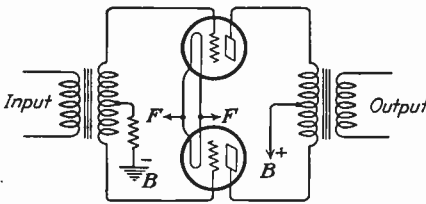


FIG. 91.—Push-pull or push-push method of connecting an amplifier.

proportional to the square of the grid-excitation voltage and consequently causes the output to resemble a linear characteristic. Class B amplifiers are, therefore, frequently classified as linear amplifiers. The linear output may readily be visualized by the graph illustration in Fig. 89. Class B

amplifiers give medium efficiency and output, while their ratio of power amplification is low.

85. The class C amplifier is practically the same as the class B, with the exception that the negative grid-bias adjustment is beyond the cut-off point of the I_p-E_c characteristic curve, generally between one-and-one-half to twice the cut-off bias voltage.

The class C amplifiers must receive a considerably greater grid excitation to overcome the heavy negative bias if suitable plate-current peaks are to be produced in the output circuit. For this reason, class C amplifiers are used only in transmitting circuits, because of the relatively higher grid voltages that may be produced in these circuits.

The output of the class C amplifier varies essentially as the square of the plate voltage within limits. The ratio of power amplification for this class of amplifier is low, while the efficiency and output of the plate circuit and output are high.

86. Balance Circuits. *a. General Considerations.*—It will later be shown that a tube circuit may be set into oscillation if a suitable coupling is provided between the plate and grid circuit so that an e.m.f. may be developed across the grid and filament (input circuit) of the tube in

question. This condition may develop either through an inductively or a capacitively reactive relationship, as in the Hartley and the tuned-plate tuned-grid oscillators, respectively.¹ The tube which is to function as an amplifier in either a transmitting or a receiving system *must under no circumstances develop sustained oscillations.*

If a receiving amplifier should go into self-oscillation, serious distortion and squealing would result owing to the effect of the oscillations "beating" against the incoming signal frequency. In addition the circuit tuning would become extremely sharp because of the decrease in the circuit losses introduced by self-oscillation. In other words, the circuit would be so selective that the band frequencies of telephone signals would be cut off so that only a narrow band of the modulated frequency would pass through, thus resulting in serious distortion.

If self-oscillation develops in a transmitting radio-frequency amplifier circuit where the plate voltages are considerably higher, the self-generated oscillations will seriously overload the tube, and the circuits to which it is connected, and may possibly ruin the tube by serious overheating. It may also burn out the radio-frequency choke coil, or the associated circuit resistances, or the plate milliammeter. When a power-amplifier tube goes into self-oscillation, the plate current in that circuit may increase to such large proportions that damage may result to the tube or circuit if they are not suitably protected with fuses or circuit-breakers.

Great care must be exercised in neutralizing or balancing a radio-frequency amplifier not only to insure protection against overloading but also to obtain the maximum circuit stability.

All radio-frequency amplifiers using the triode tube must be neutralized owing to the internal grid-plate tube capacity which enables the energy to be fed back from the plate to the grid circuit. Tubes of the screen-grid variety do not generally require neutralizing because of their extremely small grid-plate tube capacity.

b. The Hazeltine Balance.—Figure 92(a) illustrates the fundamental principle of the well-known Hazeltine system of capacity neutralization, wherein the effect of the grid-to-plate capacity is balanced by means of an output transformer.

The units L_1C_1 represent a typical radio-frequency tuning system which may be adjusted to any frequency by the variation of L_1 or C_1 , usually the latter. Let us assume the condenser C_1 to be adjusted to its minimum capacity so that the inductance L_1 is the predominant frequency factor. That is to say, the frequency of the circuit is now dependent upon the physical period of the inductance alone. It will be remembered that when the capacity across an inductance is increased,

¹ If the reader has no previous knowledge of the conditions which produce self-oscillation, it is recommended that he study Sec. 92 on General Oscillator Considerations before proceeding with this section.

the frequency of the circuit is decreased; or, in other words, the wave length increases with an increase of C across L . Since L_2 , represents an untuned primary winding of a radio-frequency transformer, it can be seen that the physical period of this winding can be made to be at or near the physical period of inductance L_1 when the condenser C_1 is at a minimum capacity setting. Then the circuit is more apt to oscillate at this point than at any other owing to the resonant conditions existing between L_1 and L_2 . Consequently, any potential across L_2 , when it is in resonance with L_1 , will induce an e.m.f. across L_1 through the tube capacitance C_t which may produce an oscillating condition.

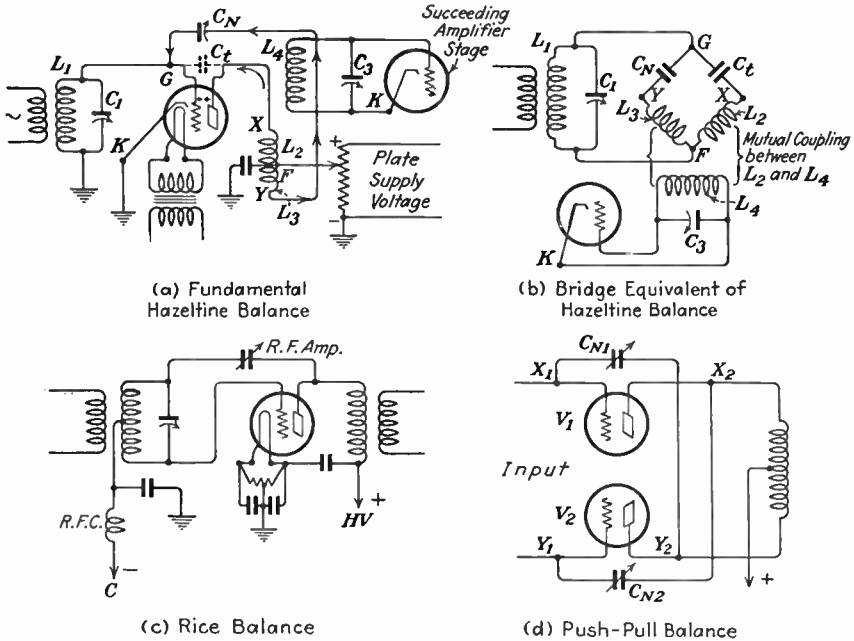


FIG. 92.—Balance or neutralizing circuits.

It will be noted from the circuit diagram (Fig. 92a) that the direction of the feed-back e.m.f. will be from the plate to the grid through the grid-plate tube capacity. This is due to the fact that the potential built up at the free end of a coil is considerably greater than at the connected end. In this circuit the end of the coil connected to the plate of the tube is considered the free or high-potential end of the coil. Hazeltine has shown that by the introduction of a third coil, or an equivalent arrangement, feed back to the grid from the plate may be overcome through the medium of an e.m.f. opposite and equal to that of the feed-back e.m.f. Referring to the circuit diagram (Fig. 92a) it will be seen that the high-potential ends of L_2 and L_3 are at the points X and Y, respectively. Thus,

if the potentials X and Y are equal, by arranging them so that they oppose each other in the tube capacitance C_t , the feed-back effect from the plate to the grid will be canceled and the tube is said to be neutralized. This is done by connecting point Y to the grid through the condenser C_N which is called the *neutralizing condenser* and which serves to control the value of the opposing e.m.f. by virtue of its reactance.

The following explanation presents a more theoretical analysis of tube-capacity neutralization.

Figure 92(b) illustrates what is known as an alternating-current bridge-circuit arrangement and is equivalent in principle to the Hazeltine method of neutralization as applied to the circuit (a). The circuit illustrated differs from the conventional alternating-current bridge because of the magnetic coupling effect between L_2 and L_3 . That is to say, there is mutual coupling between the two arms of the bridge, and this value of mutual inductance due to the coupling must be added to the value of L_2 and L_3 . Furthermore, the inductive relationship of the secondary winding L_4 to L_2L_3 must also be taken into consideration because it will affect their normal values. This condition of coupling reaction between the secondary and the two halves of the primary is usually taken care of by keeping the inductance of L_2 and L_3 nearly equal and using very tight coupling. This will cause the two circuits to affect each other equally.

Hence, if the bridge is properly balanced, the ratio is

$$\frac{C_N}{C_t} = \frac{L_2}{L_3} \quad (105)$$

This formula holds true provided the values of L_2 and L_3 include the mutual inductance of one another and also that of the secondary coil L_4 .

Consequently, if L_2 and L_3 have their turns wound in the proper directions, their respective magnetic fields will be in opposition, and, therefore, no e.m.fs. will be induced into the secondary circuit C_3L_4 . In other words, no e.m.fs. will be developed between the points X and Y owing to the opposing currents flowing through L_2 and L_3 .

On the other hand, when a high-frequency current is flowing in the secondary circuit L_4C_3 , a varying magnetic field will be produced about L_4 which will produce a difference of potential (e.m.fs.) across the coils L_2 and L_3 . These e.m.fs. will be in opposition owing to the bucking effect of the windings on L_2 and L_3 and will, therefore, cancel out. This presumes, of course, that the capacity C_N is adjusted so that both arms of the bridge balance in respect to impedance value. This equivalent circuit (b) should be compared with the schematic diagram (a). The lettering on both diagrams are equivalent for the points and circuit values indicated. When no current flows in either arm of the bridge, it is neutralized.

c. The Rice Balance.—Whereas the Hazeltine balance neutralizes the grid-to-plate capacity by the application to the grid of reverse e.m.fs. taken from the output transformer of the tube circuit, the Rice balance accomplishes the same purpose by taking the neutralizing e.m.fs. from the input transformer and applying them to the plate. The scheme is illustrated in Fig. 92(c).

The Rice balance is very successfully used in many Western Electric transmitters employing single triodes as radio-frequency amplifiers.

d. The Push-pull Balance.—This system of neutralization is employed with push-pull radio-frequency amplifiers and is shown in Fig. 92(d). Assuming that the neutralizing condensers C_{N_1} and C_{N_2} are not present, it is seen that e.m.fs. of opposite polarities would be fed back to the grids of both tubes through the tube capacities. Assume that the plate of V_1 feeds back a positive (+) potential to its grid and V_2 feeds back a negative (-) potential to its grid. Now, if the neutralizing condensers are connected as shown, it is evident that the negative potential fed to the grid of V_1 from the plate of V_2 , through the neutralizing condenser C_{N_1} , will neutralize any equal positive (+) feed back to the grid of the same tube from its own plate. The same holds true for the other tube which is neutralized by energy fed from the plate of V_1 through the neutralizing condenser C_{N_2} . Thus the circuit is effectively neutralized.

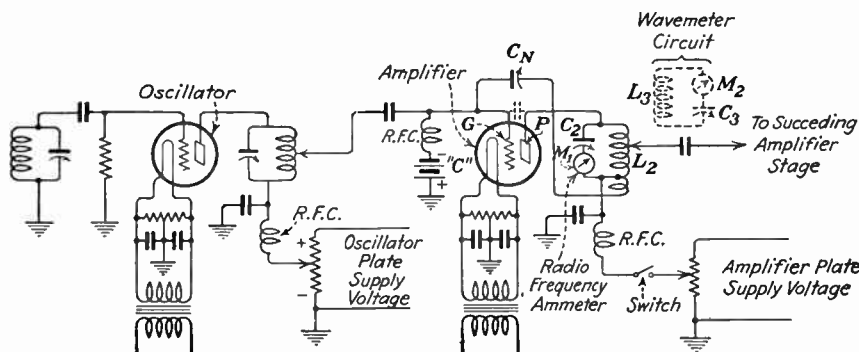


FIG. 93.—Illustrating methods of neutralizing radio-frequency amplifier in transmitting circuit.

87. Neutralization of a Transmitting Circuit.—In the neutralization of a transmitting radio-frequency amplifier circuit some visible indicating method must be used in order to determine if the amplifier is properly balanced.

Figure 93 illustrates a fundamental arrangement of a transmitting radio-frequency amplifier, showing the oscillator (driver) and the amplifier. Although a number of intermediate radio-frequency amplifiers may be used, only one stage is shown to illustrate the fundamental procedure which follows.

1. The plate-voltage-supply line on the amplifier stage must first be opened so that no potential whatever is applied to the tube. This is accomplished by opening the plate-supply switch.

2. The oscillator is then set into operation by closing the plate- and filament-supply circuits and setting the tuning adjustments to the desired operating frequency.

3. The oscillations generated by the driver tube are now applied across the grid and filament of the amplifier tube. Although this tube will not amplify the generated oscillations owing to the open plate-voltage circuit, a radio-frequency current of small

amplitude will circulate in the oscillatory circuit L_2C_2 owing to the coupling of the internal tube capacity GP , if the neutralizing condenser C_N is at minimum capacity.

4. If no current indication is obtained on the radio-frequency ammeter M_1 , the circuit C_2L_2 may be improperly adjusted and the condenser C_2 should be varied. If there is still no indication, a more sensitive thermocoupled radio-frequency ammeter must be inserted to record the small radio-frequency currents circulating in L_2C_2 at the resonant adjustment.

5. When an indication is obtained, vary the neutralizing condenser C_N until the oscillatory mesh current in L_2C_2 drops to minimum. The tube capacitance is now neutralized and the plate-supply voltage may be applied to the amplifier tube by closing the plate-supply switch.

The same procedure may be duplicated through any number of radio-frequency amplifier stages but all the circuits must be carefully retuned to insure the maximum resonant conditions between the circuits.

Another simple method for ascertaining the mesh currents in the radio-frequency stage to be neutralized is to place a standard wavemeter near the plate coil L_2 and adjust it to resonance with the driver frequency. When the circuit L_2C_2 is properly adjusted to this frequency the small mesh currents that flow in this circuit may be picked up by the wavemeter circuit L_3C_3 and indicated by the thermogalvanometer M_2 . The neutralizing condenser C_N may then be varied until this reading is reduced to zero, thus indicating that the circuit is neutralized. Care must be exercised not to bring the wavemeter too close to the mesh circuit, to prevent the possibility of damage to the sensitive thermogalvanometer M_2 .

This method cannot be used when the oscillatory mesh L_2C_2 is enclosed in a shielded compartment owing to the reduction of the exterior magnetic field around the compartment.

Balanced radio-frequency amplifiers, as illustrated in Fig. 92(d), are neutralized in exactly the same manner as for single tubes. It is important, however, that both of the neutralizing condensers have the same capacity and that all leads connected from the plate and grid circuits of both tubes are of the same lengths.

In high-quality transmitter design the two rotary plates of the neutralizing condensers are simultaneously operated by gears in order to obtain the exact capacity requirement and at the same time to make possible the neutralization of both tubes with a single control.

88. Magnetic Shielding.—The stability of radio-frequency amplifiers in transmitting and receiving depends primarily upon the neutralization of the tube capacity and the magnetic shielding of the associated tube circuits. If a radio-frequency amplifier is properly neutralized, self-oscillation may still occur if reactions take place between the grid and plate circuits owing to the magnetic coupling of coils or associated leads.

Figure 94 illustrates the arrangement of magnetic shielding in a radio-frequency amplifier circuit. The magnetic fields that are produced

around the radio-frequency coils L_1L_2 and L_3L_4 , when an alternating signal e.m.f. is applied to the grid of the preceding tubes or exciting circuit, are confined almost entirely to the enclosed compartments of the shields 1 and 2, as illustrated. The ability of the shielding to maintain the magnetic fields to the confines of the compartment depends upon the construction of the entire unit. This may be clearly seen if the theoretical functions of the circuit and shields are analyzed.

When a varying magnetic field is moving about the radio-frequency coils, the flux will tend to spread out in a more or less circular movement in which the distance of its travel depends upon the intensity of the currents flowing in the respective coil circuits. If, therefore, the fields

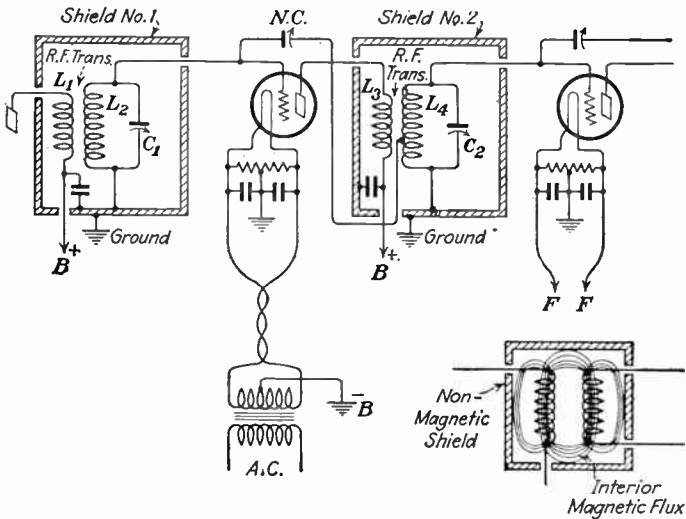


FIG. 94.—Illustrating effect of magnetic shielding on radio-frequency circuits.

are strong enough, and the two radio-frequency transformers are in the magnetic axis of each other, magnetic reactions will take place resulting in self-oscillation and instability. If, however, the coils are enclosed in a metallic compartment of non-magnetic material, such as copper, aluminum, or brass, and assuming that the compartment forms a complete electrical circuit, the varying magnetic fields around the coils will tend to displace electrons in the shield at the particular frequency in which the magnetic field is varying. This action generates small magnetic whirls, or eddy currents, which in turn produce their own magnetic fields and oppose the field created by the high-frequency currents circulating in the coils. This opposing effect varies directly with the frequency so that, as the frequency increases, the shielding effect is greatly improved. In other words, if the shielding material is made of a high-conducting metal such as copper, and thoroughly closed to form a low-resistance circuit, the

intensity of the magnetic field created by the eddy currents in the shield will oppose the coil fields and thereby reduce the external effect of the magnetic fields upon other coils or wires.

The efficiency of the shielding depends a great deal upon the high conductivity of the material used, because the greater the current that is flowing, the more effective will be the opposing or shielding effect. Great care must be exercised, therefore, that all ends of the compartment are thoroughly soldered to form a complete current path.

89. Inductance and Capacitive Changes with Shielding.—The inductance of a coil will decrease when enclosed in a non-magnetic shield owing to the reduction of the magnetic flux around the coil. This decrease in the inductance takes place because the magnetic fields set up by the eddy currents in the shield circuit are in opposition to the coil flux resulting in a total reduction of the lines of force. Hence, since the inductance of a coil depends upon the amount of energy that is stored up in magnetic form, any decrease in the number of the magnetic lines results effectively in a decrease of its inductance.

The electrostatic capacity, however, increases when a coil is placed into a non-magnetic shield owing to the relationship of the various parts of the coil to the shield. This condition is illustrated in Fig. 95. The capacity effect may greatly increase the effective resistance of the coil, particularly at high radio frequencies, since the condenser effect introduces additional energy dissipation or loss. The losses, due to shielding, may be considerably reduced, however, by using a shielded compartment of large dimensions so that the high-frequency resistance losses are minimized. Close-fitting shields must be avoided, if the maximum degree of efficiency (high Q) of a coil is to be obtained. The shield should be equidistant from all portions of the coil with a spacing of at least one-half of the diameter of the coil.

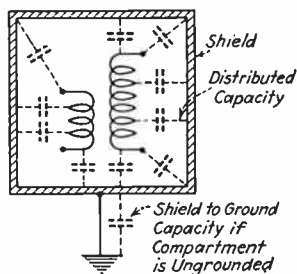


FIG. 95.—Capacity effects between coils and shielding.

The shields must be thoroughly grounded to avoid a "floating" effect of the shield above the ground potential. That is to say, if the shielding is not connected to a ground, an electric field (a capacity) may exist between the shield and the ground owing to the potential difference between these two points. If, therefore, any wires or circuit parts are situated in this field, serious instability of the entire circuit may result owing to the effect of the field upon associated circuits. In some circuits, for example, particularly at very high radio frequencies, the connecting of two shielded compartments to the ground, as illustrated in Fig. 94, may produce considerable instability, particularly if the two compartments are separated some distance from each other. It would seem that

if the two compartments were connected to the same ground on a metal chassis, no difference of potential would exist between these two points since they appear to be at the same potential. At very high frequencies, however, this is not true because the distance of several inches may introduce a voltage drop due to the resistance introduced at the high frequencies. In other words, the resistance between several inches in a metal chassis may be zero as far as low frequency or direct currents are concerned but may be fairly large at high frequencies.

At very high frequencies all common return leads must be connected to one point in as direct a manner as possible to reduce the voltage drops along the leads. If all leads are to be connected to a common chassis then a heavy strip of copper ribbon several inches wide should be used to reduce the voltage drops at the very high frequencies.

Great care must be taken that grid and plate leads are made as short as possible so that no magnetic or electrostatic fields may exist between these leads with preceding or succeeding radio-frequency stages. At extremely high radio frequencies, or in any high-gain amplifier, the grid and plate leads that extend outside of the shielding compartment should also be thoroughly shielded by a metal sheathing and grounded. Even the tubes themselves should be completely enclosed in a metal casing, particularly those of the high-gain screen-grid variety. Complete shielding of all high-frequency units is imperative if the maximum degree of efficiency and stability is to be obtained.

Low-frequency amplifiers and associated apparatus, such as audio-frequency transformers, power transformers, and choke coils, are generally shielded with magnetic materials to reduce the effect of the exterior flux upon near-by circuits. A magnetic shield of low reluctance provides a better path to the magnetic flux so that the exterior effect outside of the shielding is negligible. In effect, the magnetic shield acts as a short circuit to the flux path and thereby nullifies the magnetic field outside of the shielding.

90. Grid-biasing Methods.—It has been shown that the proper operation of all vacuum-tube circuits depends primarily upon the adjustment of the negative grid voltages or grid bias. Any one of many methods may be used to procure a definite value of grid bias. These are illustrated in Fig. 96.

Circuit (a) is the fundamental method of obtaining a C bias by the use of a cell or battery connected between the filament and the grid. The e.m.f. established across the cell terminals is applied directly across the grid and filament of the vacuum tube at the points E_c .

Circuit (b) is the same as circuit (a) with the exception that the negative C bias is supplied by a direct-current generator. This method is used only in transmitting circuits where high values of grid bias are required.

Circuit (c) is the most common system used in transmitting circuits where all-alternating-current operation is required. Here both the plate and grid-bias voltage supply is obtained from a rectifier and voltage-divider system. When using this method great care must be exercised in filtering the plate and grid-bias supply, particularly the latter. If the grid-bias filtering permits the slightest variation in voltage so that a ripple is produced, the amplifying property of the grid will cause the

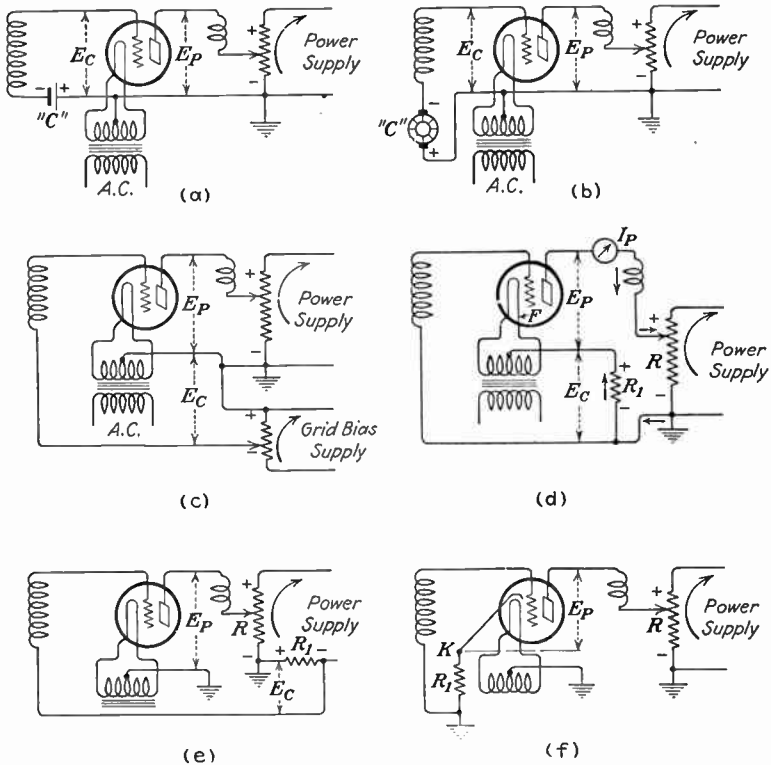


FIG. 96.—Methods of procuring grid bias.

ripple frequency to modulate the steady-flowing plate current resulting in serious distortion and hum. Typical plate and bias power-supply filtering methods are fully explained and illustrated in Chap. XIII on Rectifier Units.

In circuit (d) the grid bias is obtained as follows: when the filament is lighted and a steady direct-current potential is applied at E_p owing to the e.m.f. supplied by the voltage divider (connected across the power supply), a steady electron flow from the filament to the plate will take place in the direction FPR_1F . This constitutes the normal flow of plate current I_p , as indicated by the plate milliammeter.

Now, since the electrons flowing in the plate circuit must return to the filament to complete the circuit, any resistance connected between the negative terminal of the voltage divider and the filament will have an e.m.f. developed across it of a value $E = IR$. Thus, if the plate-current flow for some fixed value of plate potential is 20 ma. or 0.02 amp., and the value of the resistance R_1 is 2,000 ohms, an e.m.f. of 40 volts ($E = IR$) will be developed across the resistance. Hence, since the negative end of the resistance is connected to the grid, and the positive end to the filament, this voltage will be applied between the grid and the filament of the tube. In other words, the grid is 40 volts negative with respect to the filament owing to the voltage developed across the resistance.

Circuit (e) illustrates the same principle, with the exception that the biasing resistance R_1 is connected in the negative lead of the plate-supply system. Thus, the voltage developed across R_1 will depend upon the value of the resistance and the current flowing in the power-supply circuit. This is the common biasing method used in receiving circuits where the filament type of tube is used, such as in the final power-amplifier stage.

Circuit (f) is the biasing method used with the cathode-heater type of tubes. The operation is identical with that of circuit (d), with the exception that the biasing resistance is connected in series with the cathode instead of the filament.

When using the circuits shown in (d) and (e), the plate voltage is that of the power supply less the bias voltage. It is apparent, therefore, that to supply, for example, 180 volts to the plate and 33 volts negative bias, the power supply must provide 213 volts.

In actual practice all biasing resistances must be shunted by a condenser in order to provide a low-reactance path for the signal currents during the period in which the tube is receiving incoming signals. If a condenser is not provided, considerable signal energy will be dissipated in the biasing resistor. The proper arrangement of the biasing resistance and the by-pass condenser is shown in Fig. 84.

91. Radio-frequency Chokes.—*a.* The radio-frequency choke coil is an inductance used in transmitting and receiving circuits to keep the high frequencies out of the plate-power and grid-voltage supply circuits. The proper isolation of radio frequencies from supply circuits by the use of chokes and by-pass condensers also provides greater circuit efficiency, since the radio frequencies will be forced to take the path of lower reactance through the by-pass condenser connected ahead of the choke. Furthermore, the isolation of these radio frequencies from the supply circuits also prevents reaction between circuits which are connected to a common power-supply system thus insuring a maximum degree of circuit stability. Circuits which do not provide proper isolation because of choke coils, or resistors in the grid and plate circuits and receive their

power supply from the same source react with each other owing to what is generally known as *common impedance coupling*.

When choke coils are not used it is also possible to prevent reactions due to this coupling by the use of grid- and plate-filter resistances as previously explained. The disadvantage of the resistance-isolation method particularly in high-current-carrying circuits is the IR drop produced across the resistances. Although this method is highly efficient and generally used in low-current-carrying circuits such as in receivers, the choke-coil method is preferred where a high-impedance isolation is required with a minimum of direct-current resistance or IR drop.

The design of the choke coil however, to perform efficiently as a high-impedance reactor is relatively a much more difficult problem since

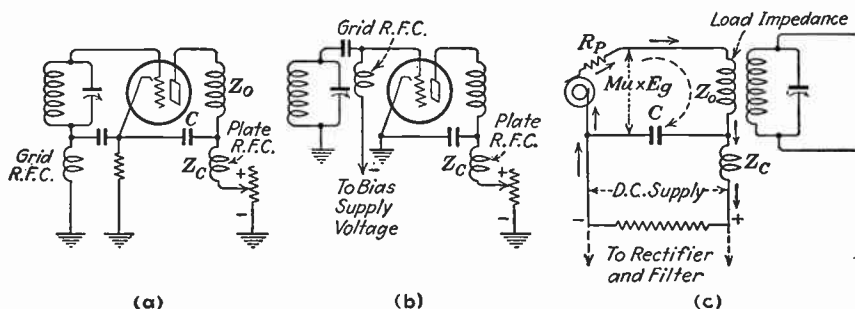


Fig. 97.—Common methods of connecting choke coils in plate and grid circuits.

the distributed capacity X_c between the turns in the winding may provide a low-reactance path around the high coil reactance X_L . That is to say, the frequencies instead of being impeded are actually by-passed through the capacity of the coil thus making the isolation desired ineffective.

Figure 97(a) and (b) illustrates the most common arrangement of radio-frequency choke coils (R.F.C.) in the plate and grid circuits. Diagram (c) represents the theoretical circuit which may be analyzed as follows.

During the period in which no e.m.f. E_g is developed, a steady plate current will flow through the circuit in the direction of the straight arrows. This flow will be unaffected by the impedances Z_o and Z_c since it is assumed that a well-filtered direct-current has been developed at the power-supply source. Small IR drops may take place across these impedances, however, owing to the direct-current resistance present in the coil windings, but these are negligible.

Now let us assume an incoming signal of $\mu \times E_g$ to be developed in the equivalent plate circuit of Fig. 97(c), and that it is desired to keep the plate-current variation due to this signal (called the *alternating-current plate component*) out of the direct-current supply circuit. In order to accomplish this, the alternating-current plate-current component must be provided with a low-reactance path Z_oC in the direction of the dotted arrow. By the application of alternating-current engineering principles it will be seen that the total signal current flowing through Z_oC will be dependent upon

the ratio of the impedance of the choke coil Z_c to the by-pass reactance of C , thus:

$$\text{By-passing efficiency} = \frac{Z_c}{X_c} \quad \text{or} \quad \frac{R}{X_c} \quad (106)$$

Now let us assume that the lowest radio frequency to be developed in the plate circuit through $Z_c C = 500,000$ cycles and the highest frequency 1,000,000 cycles. The value of the capacity and choke coil should be so designed that the impedance to the lowest frequency through C , is extremely low, while the impedance through Z_c is very high. If the choke coil has a value of 100 mh. and a direct-current resistance of 200 ohms, then the impedance to a frequency of 500,000 cycles will be 314,000 ohms and that to the frequency of 1,000,000 cycles will be 628,000 ohms.

The percentage of current flowing through the condenser C will be Z_c/X_c . Hence, if it is desired to increase the percentage of the high-frequency component flowing through $Z_c C$, it will merely be necessary to increase the ratio to the point in which the ratio approximates nearly 100. When this has been accomplished, the reactance of Z_c will be considered sufficiently high to isolate the frequencies from the power-supply circuit. Hence, if the impedance value of Z_c is very high, a smaller capacity may be used at C . This is a theoretically desirable condition since, if the value of C is kept small, condensers with mica dielectrics may be used thereby insuring a minimum dielectric loss. Small-capacity condensers will also have less inductive effect—a condition which is generally present in larger capacities and highly undesirable in radio-frequency circuits. Unfortunately, however, it is extremely difficult to increase the inductance of the choke coil without introducing too much distributed capacity and reducing the effective impedance. Even though the inductance may have a high reactive value theoretically, the distributed capacity shunted across it will greatly reduce its effective impedance, particularly at the higher frequencies. The efficiency of the choke coil, therefore, depends greatly upon its design, and in highly efficient circuits the choke generally consists of several spaced “pie” sections, wound on an insulation form of Isolantite, connected in series to reduce the distributed capacity and to reduce dielectric absorption and losses. The size of wire used will depend upon the heat-carrying capacity of the winding, particularly in the plate filtering circuits of vacuum-tube transmitters where the plate currents may be considerable.

b. Choke-coil and Plate-filter Design.—If the ratio of the impedance Z_c to the reactance X_c (Fig. 97) is nearly 100, then the current flow through C will be a maximum. Practically, if this ratio is even as low as 60, the percentage of plate-current flow through C will be very close to maximum. As the ratio falls below 60, however, the percentage of flow through C decreases relatively more rapidly. A value of capacity should be used in which the ratio is equal to at least 60. Assume the value of C to be 0.5 mf. and the frequency 500 kilocycles, giving an X_c of 0.6 ohms. Now, if a radio-frequency choke of 20 μ h. is inserted at Z_c it will have a reactance of approximately 63 ohms at a frequency of 500,000 cycles. Then the ratio of

$$\frac{Z_c}{X_c} = \frac{63}{0.6} = 105$$

This would indicate, in accordance with the above explanation, that the ratio is sufficient to pass through C a maximum value of the alternating-current plate-current flow. This would be true if the reactance due to Z_c were pure inductance containing absolutely no distributed capacity, but since this is practically impossible the value of Z_c must be made considerably larger to introduce sufficient impedance to overcome the inherent distributed capacity. Roughly the value of the choke impedance Z_c for radio-frequency circuits should be about 4,000 times greater than the ratio indicated, or 20 μ h. \times 4,000 = 80 mh. No hard and fast rules can be applied to an accurate calculation of radio-frequency chokes owing to the many variables introduced by

distributed capacity, high-frequency resistance, and design. Although it is possible to obtain a relatively greater degree of accuracy by the use of resistances in place of choke coils, these, too, present difficulties due to their IR -drop losses, inductance, and distributed capacities, particularly at the higher radio frequencies.

In low-frequency amplifiers, where resistances are used in place of Z_c , this ratio is highly accurate since the distributed capacity is a negligible quantity at these frequencies. Hence, if the circuit in Fig. 97(c) is a low-frequency amplifier circuit in which Z_c is the primary winding of an audio-frequency transformer and Z_c is replaced by a resistance R , and if C is a low-frequency by-pass condenser, then the values of C and R may be readily computed by the ratio R/X_c . For example, if the ratio of 100 is assumed and a frequency of 50 cycles is applied, then the values of R and C to effect efficient by-passing and isolation from the power-supply circuits must be 300,000 ohms and 1 mf., respectively. Owing to the fact, however, that there is very little difference in the filtering efficiency between a ratio of 100 or 50 in audio-frequency systems, the latter is preferable since it will permit the use of a lower plate-filter resistance and thereby reduce the IR drop across the resistance to the plate of the tube. Thus, at a frequency of 50 cycles, a plate-filter resistance of 150,000 ohms may be used, since $R/X_c = 150,000/3,000 = 50$. The values of R and C are always calculated on the basis of the lowest frequencies that are to be amplified, since the reactance of C decreases with an increase of frequency.

If lower resistances are desired to effect lower IR drops, the value of C may be increased proportionately to produce the same result. In this case a condenser value of 10 mf. having a reactance of 300 ohms at 50 cycles and a plate-filtering resistance of 15,000 ohms would produce the same result.

In receiving circuits the resistance method of radio-frequency filtering is advisable because in the low-current-carrying circuits the IR drop is negligible. The resistances used for this purpose should be of the non-inductive type.

III. VACUUM-TUBE OSCILLATORS

92. General Oscillator Considerations.—A vacuum-tube oscillator consists essentially of an oscillatory circuit [generally a parallel-resonant circuit Fig. 99(b)], a vacuum-tube amplifier, and a feed-back circuit. In the oscillatory, or output, circuit, electrical oscillations occur according to the fundamental laws governing this type of operation. The feed-back circuit takes a part of the energy from the output circuit and impresses it in the form of a voltage on the grid and filament (the input circuit) of the vacuum tube, where it is amplified and again returned by the plate-filament circuit to the output circuit, supplying the losses occurring in the latter and thus keeping it oscillating. The feed-back circuit may consist of capacitive coupling through the grid-plate capacity of the tube, or inductive coupling through a coil in the grid circuit magnetically coupled to the output circuit. Thus, the vacuum tube itself does not oscillate but merely amplifies the electrical impulses impressed on its grid from the oscillatory circuit and delivers this amplified energy back to the output circuit to make up for the losses therein, in order to maintain its oscillations.

93. Fundamental Oscillator.—The fundamental circuit in Fig. 98 shows a simple vacuum-tube oscillator capable of producing waves of a

continuous character. The theoretical functioning of this circuit is identical to almost every form of tube transmitting-circuit oscillator and should be carefully analyzed.

The moment the filament switch is closed and the direct current is applied, the filament emits electrons which bombard the plate. A plate current will then flow from *F* to *P* through coil L_1 back to the negative terminal of the filament, completing the circuit and resulting in a steady flow of current in the plate circuit.

This would indicate that a magnetic field must be present about coil L_1 . The field would appear to be steady because the flow of plate current is assumed to be constant. The moment the switch was closed, however, the magnetic field about the coil L_1 expanded, which, according to the electromagnetic law, would cut the coil L_2 in a given direction owing to its inductive relation and would, consequently, induce an e.m.f., or difference of potential of a given polarity across it. If, then, the coils are wound in a direction in which L_2 would receive a positive potential at the top, and a negative potential at the bottom, the grid which is connected to the top of the coil

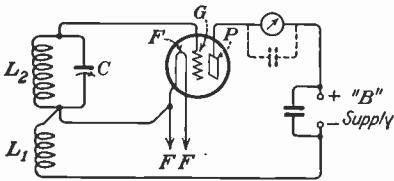


FIG. 98.—Fundamental oscillator circuit.

would receive a positive charge. This partly neutralizes the space charge between *F* and *P* immediately and allows more current to flow in the plate circuit PL_1F . This results in a greater expansion and movement of the field in the same direction about L_1 and consequently a greater e.m.f. of the same polarity is induced into L_2 , resulting in a greater charge upon the grid. Again the plate current increases, and a heavier charge will

again be placed upon the grid. This action will continue up to a certain point, depending upon the characteristics of the tube and the resistance of the circuit. As soon as the plate current ceases to increase, the field around L_1 collapses and in so doing cuts L_2 in the opposite direction inducing an e.m.f. in L_2 of a reversed polarity. Thus, as the increase of the field around L_1 and L_2 made the grid positive, a decrease in the plate current will result in making the grid negative. Thus, if the positive charge on the grid increases the plate current, it is quite apparent that a negatively charged grid will decrease this current. The plate current will, when the grid is negative, decrease to a point below normal, just as it will increase above normal when the grid is positive. Hence, the plate current will decrease to a point at which there will be no further change in the grid potential. The complete cycle will then be reversed, and the operation will be completed all over-again. Thus the plate current will rise and fall with a definite frequency, the period of which depends upon the values of the inductance and capacity L_2C in the circuit. By using the proper constants, a circuit of this type can be arranged to produce frequencies ranging from the lowest audible range to the highest range in radio frequencies.

In all forms of oscillating systems, especially those in which the frequencies are high, precaution must be taken to keep the resistances of the constants low. Otherwise the oscillations are apt to stop, and difficulty might be encountered in bringing about consistent transmission. These losses, especially at the higher frequencies, are due to various conditions such as high-frequency resistance losses, dielectric absorption due to poor insulating materials, eddy currents, poor connections, and other loss-producing conditions.

94. The Hartley Oscillator.—Figure 99 illustrates the circuit arrangement of the well-known Hartley oscillator. The principles of oscillation are the same as in the fundamental oscillator described in the preceding section. The series-feed arrangement in diagram (a) is in fact theoretically identical with that of the fundamental oscillator.

The grid excitation takes place across the grid coil and the points A and B owing to the reactance drop X_L of the coil. The proper value of this reactance will depend upon the type of tube used and the grid-excitation voltage required. The grid excitation in both the series- and the

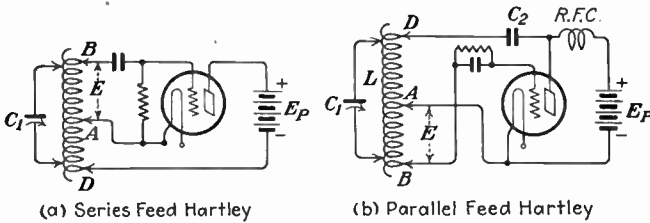


FIG. 99.—Hartley oscillator circuits.

parallel-feed arrangements takes place through the medium of the induced c.m.f. from the plate portion AD of the coil to the grid portion AB with the exception that in the *parallel feed* (b) the initial excitation takes place through the plate-blocking condenser C_2 in the form of a dielectric displacement of electrons. This displacement causes a movement of electrons in the plate coil which produces a magnetic field. This condenser also serves to block out the direct-current component of plate-current flow from the oscillatory circuit thereby allowing only the high-frequency currents to flow in the L and C circuits. This condenser must have a low reactance (X_c) to the operating frequencies to minimize the drop across it, so that it may deliver the proper potential across the plate-excitation coil. This condenser must, however, have a sufficiently strong dielectric to withstand the constant direct-current and oscillating potentials across it.

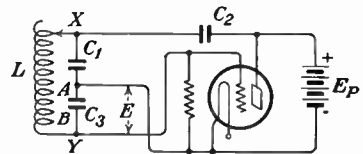


FIG. 100.—Colpitts oscillator circuit.

95. The Colpitts Oscillator.—Figure 100 illustrates the circuit arrangement of the Colpitts capacitive feedback oscillator.

The theoretical operation of this circuit is identical with that of the Hartley parallel-feed system with the exception that the grid-excitation voltage takes place across the capacitive reactance C_3 . The initial excitation is produced as before through the plate-blocking condenser C_2 in the form of an electron displacement which in turn also produces a displacement in the plate-excitation condenser C_1 . This results in a

potential difference across C_1 and C_3 which excites the grid and produces sustained oscillations.

The frequency of these oscillations will depend upon the oscillatory circuit constants L and C , thus,

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{90}$$

The total reactive voltage present across the two condensers at the point XY may be divided by the proper choice of the plate- and grid-excitation condensers C_1 and C_3 so that the correct voltage drop across C_3 may be obtained for proper grid excitation. The higher the reactance X_c of the grid-excitation condenser, the greater the potential difference or e.m.f. across it.

96. The Push-pull Oscillator.—The theoretical operation of the push-pull oscillator is identical with that of the push-pull or balanced

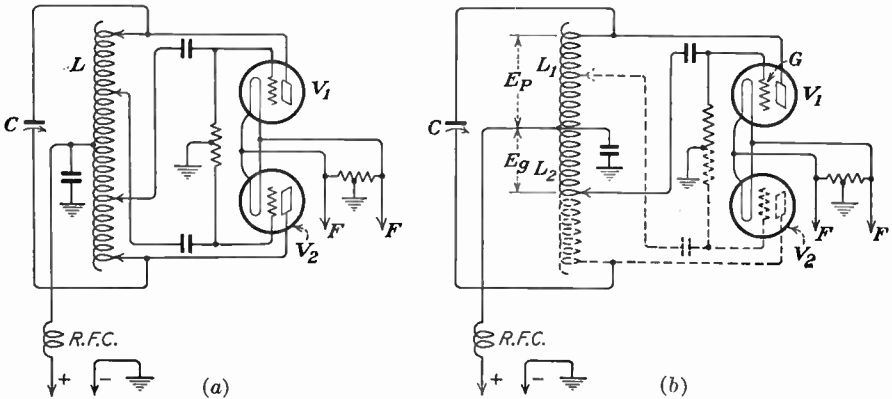


FIG. 101.—Balanced or push-pull oscillator circuit.

amplifier to be explained in a later section, with the exception that the grid voltages are developed in the circuit itself, as in any form of oscillator, through a medium of inductive or capacitive feed-back reaction.

A typical circuit arrangement of a push-pull oscillator is illustrated in Fig. 101(a). The instantaneous operation of the tube V_1 in diagram (b) now to be explained is through the circuit shown by the solid lines. At the instant of any instability in the plate or filament circuit, such as would be caused by closing the power-supply circuit, a potential difference E_p is developed across the plate-excitation coil L_1 . During the period in which the instability took place a momentary moving of the magnetic field was developed across L_1 . In accordance with the laws of induction, therefore, an e.m.f. E_g , will be developed across L_2 . This in turn excites the grid G of V_1 . Since C and L_1 represent a tuned circuit connected across the tube, the circuit will be set into self-oscillation at a frequency

depending upon the L - C constants of the circuits. After the initial impact to the oscillatory circuit has occurred, sustained oscillation will take place resulting in an excitation of both grids, 180 degrees out of phase with each other. By the proper adjustment of the two grid connections to the inductance, perfect symmetry or balance may be obtained.

The advantage derived from an arrangement of this kind is greater power output and better frequency stability owing to the reduction of the tube input and output capacity, since, with this arrangement, both tubes are effectively in series across the tuned-load circuit. This gradually reduces the circuit losses due to tube capacity and makes this form of oscillation highly desirable for use at high radio frequencies where large power-oscillation outputs are required with good frequency stability.

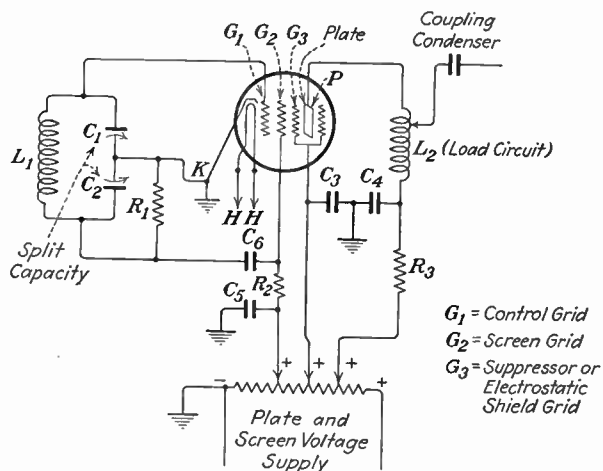


FIG. 102.—Electron-coupled oscillator circuit.

97. Conventional Electron-coupled Oscillator.¹—Figure 102 illustrates an electron-coupled oscillator which may be used in all radio-frequency circuits in which an oscillator is required. This circuit is particularly adaptable to receiving systems employing the superheterodyne principle.

The frequency of the oscillations generated is determined by the oscillatory circuit $L_1C_1C_2$ as in all other forms of oscillators. The circuit differs, however, from the conventional oscillator in that no capacitive,

¹ The actual operation of the electron-coupled oscillator in a modern receiver is explained in connection with the RCA AVR-5A receiver in Chap. XI on Radio Receivers. The following articles are also recommended to the reader desiring full information on the development of this circuit:

Dow, Electron-coupled Oscillator Circuits, QST, January, 1932;

Dow, A Recent Development in Vacuum Tube Oscillator Circuits, *Proc. I.R.E.*, December, 1931.

inductive, or direct coupling exists between the load and grid circuits. Although it is apparent that oscillations may be generated without inductive coupling through the medium of the internal grid-plate tube capacity, such as in the tuned-plate, tuned-grid, and crystal-controlled oscillator, these conditions do not occur in the electron-coupled type of oscillator.

In the electron-coupled oscillator the inter-electrode grid-plate capacity is either neutralized, or the plate is shielded by the use of a third screen grid. In the latter method, the addition of a third grid near the plate acts as an electrostatic screen, thus making neutralizing unnecessary. Hence, if no capacitance is present between the plate and the control grid G_1 , and no other electrostatic or magnetic relationship exists between the plate and grid circuits, any oscillations that are generated in the frequency-controlling circuit L_1, C_1 can affect only the load circuit through the medium of the electron flow between the cathode K and the plate P . Hence, the variations in the plate circuit, due to the electron pulses transmitted to the plate from the cathode, generate in the plate circuit a pulsating direct-current variation at the frequency of the generating circuit L_1, C_1 . This in turn produces an alternating e.m.f. in the plate circuit at a frequency determined by the pulses of electrons impinging on the plate.

Hence, since there are no reactions taking place between the load and frequency-generating circuits owing to electrostatic or magnetic coupling, and because the coupling is due solely to the electron stream to the plate, the names *electron-coupling* or *electron-coupled* oscillator are generally applied.

The exceptional frequency stability which may be obtained with this circuit owing to the negligible effect of the load-circuit variations upon the frequency-controlling circuit makes this form of oscillator desirable for low-power oscillating systems.

The initial starting of the oscillations is due to a voltage drop across the split section of the tuning condenser C_2 at some instant of instability, such as when the filament- or plate-supply circuit is first closed. This potential difference across C_2 sets the oscillatory circuit L_1, C_1, C_2 into oscillation at a definite frequency determined by the constants $f = \frac{1}{2\pi\sqrt{LC}}$.

During the initial oscillating pulse in the circuit L_1, C_1, C_2 , a voltage is built up across this combination. A part of this voltage is present across the split capacity C_1 which, since the grid-cathode terminals are connected directly across it, establishes an e.m.f. between G_1 and K . Hence, the alternating voltages built up across the oscillating circuit vary the grid voltages on G_1 in accordance with the frequency characteristics of the circuit, and the grid in turn controls the rate at which the electrons impinge upon the plate.

The condensers C_3, C_4, C_5 , and C_6 are radio-frequency by-passes to maintain a low-reactance path for the alternating-current variations in the load and screen circuits. The combination of C_6 and the resistance R_1 serves to apply a negative bias to the control grid to insure a maximum degree of stability during the period in which the circuit is oscillating. The resistances R_2 and R_3 are isolating units to keep the radio frequencies out of the power-supply circuits, thereby additionally insuring circuit stability. These may be in the form of either non-inductive resistances or choke coils.

The load circuit in an electron-coupled oscillator may be either reactive or resistive so that L_2 may be satisfactorily replaced with a resistance. The power developed in the load circuit will of course be dependent upon the relationship of the load unit to the alternating-current plate resistance, R_p , of the tube, as in all impedance-matching circuits.

98. The Dynatron Oscillator.—Figure 103 shows the circuit arrangement of a screen-grid tube used as a dynatron oscillator.

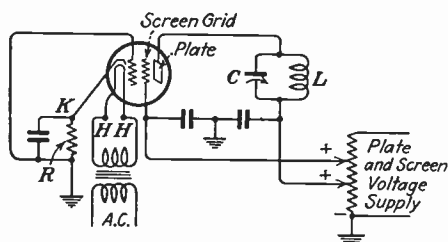


FIG. 103.—Dynatron oscillator circuit.

The dynatron oscillator presents a characteristic called *negative resistance*¹ which is well known in the theory of regeneration and which may also be used for other purposes requiring similar characteristics.

The theoretical operation of the dynatron oscillator may be described briefly as follows:

When the filament is lighted to incandescence and a plate potential is applied between the plate and the filament, an electron flow from the filament to the plate will result in accordance with the elementary theory of vacuum-tube operation.

If the grid is operated at a higher positive potential than the plate, the electrons arriving at the grid will cause an electron flow in the grid circuit of fairly large amplitude, which is, of course, true in conventional circuits when large signal potentials are applied to the grid. The important function, however, remains in the electrons that have passed through the grid laterals toward the plate. These electrons on arriving at the plate cause a breakdown of the surface tension, owing to the collision impacts, which results in an emission of electrons from the plate surface. This phenomenon is known as *secondary emission*. Now, since the grid is at a higher potential than the plate, the electrons due to secondary emission will be attracted to the grid. This presents a condition in which both the electrons from the filament and from the plate (the latter due to secondary emission) will be attracted to the grid. Now let

¹ See Regeneration, Sec. 155.

us assume the plate voltage to be gradually increased. In ordinary tube circuits an increase in plate voltage would result in an increase in the plate current, but in this circuit the gradual increase in plate voltage reduces the plate-to-filament current flow owing to the neutralizing effect at some critical point between the normal electron emission from the filament circuit and the secondary electron emission from the plate. When this particular point is reached, the plate-to-filament circuit resistance would indicate a negative resistance characteristic if an alternating-current potential were applied to the circuit. It is this peculiarity which is the basis of the action of a dynatron oscillator.

Note on Tuned-plate, Tuned-grid Oscillator: This type of circuit is identical with the crystal-controlled oscillator explained in the next chapter, except that the grid-circuit load consists of an inductance and capacity which form the grid parallel-resonant oscillatory circuit.

CHAPTER IV

TRANSMITTING-CIRCUIT PRINCIPLES

The various methods of operating a vacuum tube as a radio-frequency amplifier and as an oscillator have been described in the preceding chapter. It is the purpose of this chapter to explain the combining of these various functions into a modern transmitting circuit.

I. CONVENTIONAL CIRCUITS

99. General Transmitting Considerations.—There are certain principles which underlie the operation of *all* radio transmitters. All transmitters do not, however, incorporate *all* of these principles, the final design being determined by the service the transmitter is to perform. For example, the only difference between the modern radio-telegraph transmitter and the modern radio-telephone-transmitter circuit is that the telephone transmitter is fitted with modulating equipment, whereas the telegraph transmitter is not. In so far as the oscillator and the amplifiers are concerned, the theory of operation is the same whether the emission is a modulated or a continuous wave. Therefore, the principles explained in this chapter can be applied to all transmitters regardless of type. It is obvious, of course, that the radio-telephone transmitter utilizes all of the component parts of the transmitter here mentioned, whereas a radio-telegraph transmitter would utilize all except the modulating equipment and the audio monitor (compare Figs. 104 and 105).

100. Classification of the Radio Spectrum.—In view of the fact that the frequency on which radio circuits are to operate dominates as a

TABLE IIIA.—FREQUENCY CLASSIFICATION

Kilocycles	Wave-length, meters	Band
10— 110	3,000—30,000	Low-frequency or long-wave
100— 550	545— 3,000	Medium-frequency or medium-wave
550— 1,500	188— 545	Broadcast-frequency or broadcast
1,500— 6,000	50— 200	Medium high-frequency or medium short-wave
6,000—30,000	10— 50	High-frequency or short-wave
Above 30,000	Below 10	Very high } frequency; very short } wave Ultra-high } ultra-short }

controlling factor in their design, it is necessary to classify circuits and emissions with a common nomenclature which will at once identify the frequency range and the service in which it is used.

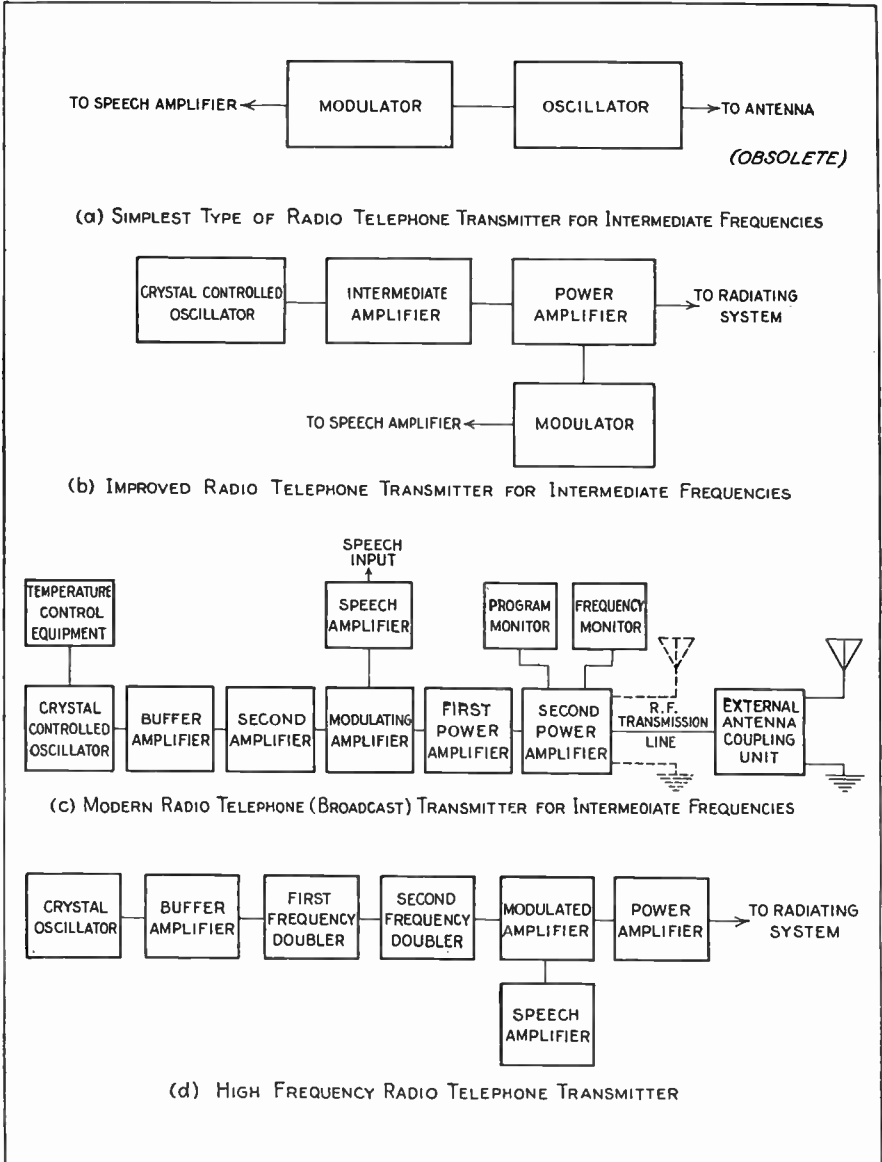


FIG. 104.—Block diagrams showing development of the radio-telephone transmitter.

The total spectrum of radio waves as shown in Fig. 106 is divided into six major bands as it appears in Table IIIA, preceding.

101. The Oscillator as a Transmitter.—The first vacuum-tube transmitters utilized the simple oscillator coupled directly to the radiating circuit as shown in Fig. 107. The circuit here shown is that of the Hartley oscillator which is typical of the simplicity of the transmitter as a whole. To increase the power of the transmitter it was only necessary to use a larger tube and increase operating voltages. Engineering conditions were ideal with the important exception that frequency stability could not be maintained, to any modern satisfactory degree, with the oscillator coupled directly to the antenna.

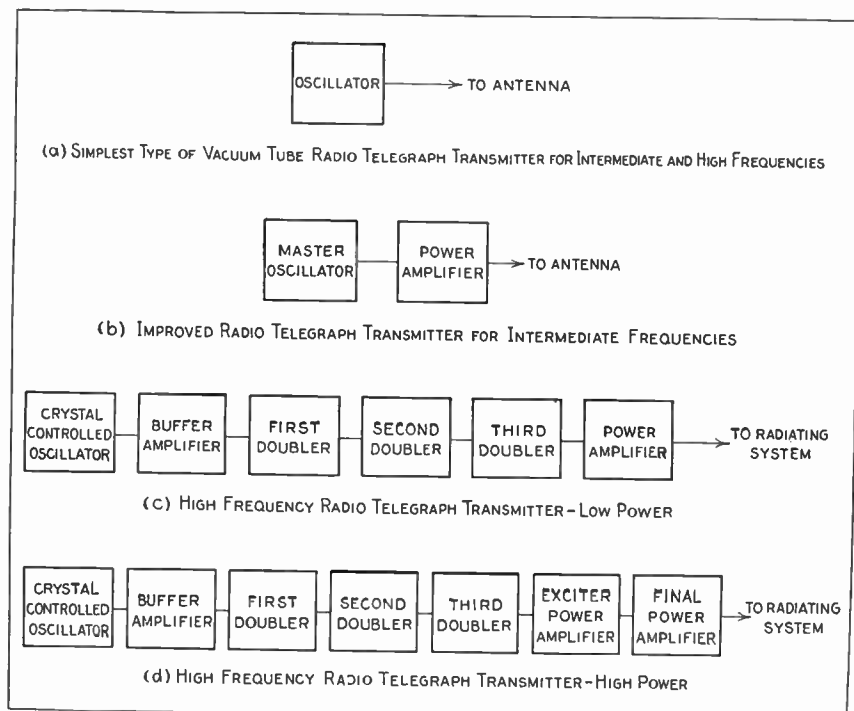


FIG. 105.—Block diagrams showing development of the radio-telegraph transmitter.

The reason for this poor frequency stability is apparent. The radiating system, that is, the antenna and ground, constituted the load on the tube which was inductively coupled to the oscillating circuit. One of the characteristics of the vacuum-tube oscillator is that its frequency of oscillation is determined by the impedance of the oscillatory circuit; if this impedance is in any way changed, the frequency is changed. Under practical operating conditions it was found that a slight change in the capacity of the open oscillatory circuit (antenna and ground) would reflect back to the oscillating circuit of the tube with a consequent change in its frequency.

This is a great disadvantage especially on board ship where the

capacity of the antenna to the ground is subject to considerable change in high winds and during movements of the ship in a sea. This circuit is now considered obsolete and is only used where economy demands simplicity and where frequency-stability regulations are not rigidly enforced. It can be tuned to any high or intermediate frequency without further modification.

102. The Oscillator-amplifier Transmitter.—It was pointed out in the preceding section that an oscillator when coupled directly to the radiating system is subject to severe frequency fluctuation. In order to overcome this deficiency the master-oscillator power-amplifier system shown in Fig. 108 was developed. This system has a decided advantage over the self-excited oscillator transmitter in that the frequency stability of the transmitter is greatly improved. This is due primarily to the fact that in this system the oscillator is not coupled directly to the antenna and, therefore, is unaffected by any changes in the antenna-to-ground capacity. The master oscillator in this particular circuit is of the Hartley type, although any of the conventional oscillators might be used with equally good results.

If the antenna-to-ground capacity does vary, the reaction upon the power-amplifier circuit will merely result in a decrease in the radiation efficiency owing to the change in the resonance conditions between these two circuits. In practice this is usually unnoticeable. The frequency stability of the transmitter is, therefore, much improved by using the master-oscillator power-amplifier system in preference to the simple oscillator alone as a transmitter.

The master-oscillator power-amplifier system is commonly referred to as the M.O.P.A. transmitter. The

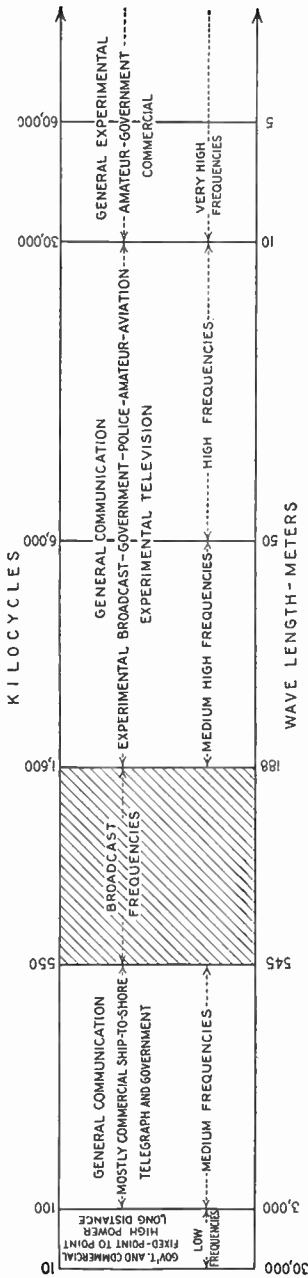


Fig. 106.—The radio-frequency spectrum and allocations.

circuit shown here can be used on any frequency to which the oscillator can be tuned.

The frequency of the oscillations in Fig. 108 is determined by the circuit constants C_1 , L_1 as before. The amount of plate and grid reactance is determined by the number of turns between the points AB and

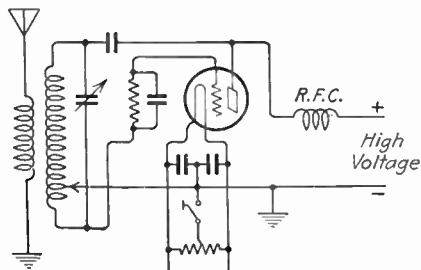


FIG. 107.—The oscillator as a complete transmitter.

AC , respectively. These reactances will govern the degree of grid-voltage excitation E_2 across the grid filament of the oscillator tube.

When the tube is oscillating at a definite frequency as determined by the constants L_1 , C_1 , the input (grid filament) to the power-amplifier tube may be connected across a portion of the inductance L_1 , so that it may

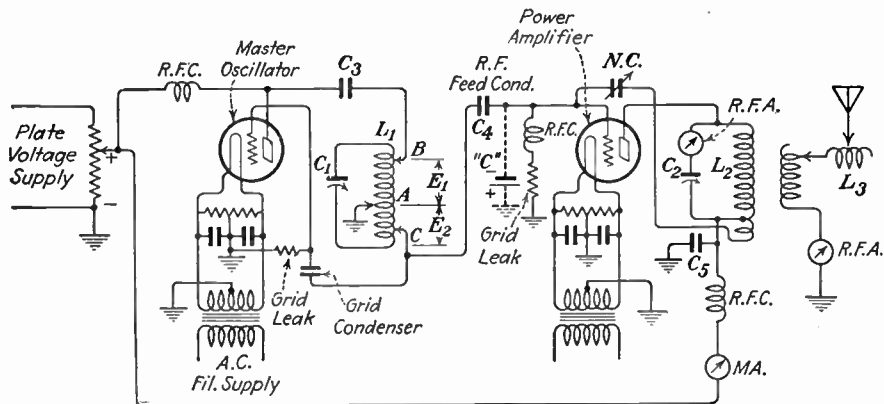


FIG. 108.—The M.O.P.A. (master-oscillator power-amplifier) transmitter.

be properly excited. In other words, the high-frequency e.m.f. built up across the grid section of the oscillator coil AC may be applied across the input to the amplifier tube through the radio-frequency feed condenser C_4 and through the common ground connection to the filaments.

The proper amount of grid excitation across the amplifier tube is determined by the reactance of the feed condenser C_4 and the point at which the amplifier grid connection is made on the oscillator coil. The power amplifier is adjusted to the proper grid and plate voltage in accord-

ance with the class of amplification in which the tube is to be operated, namely, class A, B, or C.

The grid-leak method of biasing in the power-amplifier systems is only recommended in transmitters used for telegraphy. Radio-telephone transmitters should use class B and C amplifiers to obtain a maximum linear output.

The output circuit L_2, C_2 is now adjusted to resonance with the master-oscillator by an indication of the maximum current in the radio-frequency ammeter, *RFA*, in this circuit or by a maximum dip in the reading of the plate milliammeter *MA*, after it has first been neutralized to prevent self-oscillation by properly adjusting the neutralizing condenser *NC*.

The antenna circuit is then tuned to resonance by adjusting the coupling and the tuning inductance L_3 for a maximum indication on the antenna radio-frequency ammeter.

Note: In an M.O.P.A. set, after bringing the antenna circuit to resonance, the power amplifier should again be slightly readjusted to resonance, minimum plate current. It must be realized that the antenna circuit, when tuned to resonance, should couple into the power-amplifier tank and appear as a power resistance load, but, owing to several variable factors, this coupled-in reactance will sometimes appear slightly capacitive or inductive. This means that the power-amplifier tank is not in exact resonance and, hence, the need for a slight readjustment of the power-amplifier variometer.

Condenser C_3 is the plate-blocking and radio-frequency by-pass condenser combined. Condenser C_5 is the power-amplifier radio-frequency by-pass condenser which serves as a low-reactance path for the radio frequencies in the plate circuit to flow through to the filament. This condenser in conjunction with the radio-frequency choke isolates the high-frequency currents from the power-supply source, thereby increasing the efficiency and stability of the amplifier circuit.

103. The Crystal-controlled Transmitter. *General Considerations.*—Comparing the block diagrams of Figs. 104 and 105, and considering the evolution of vacuum-tube transmitters, it at once becomes apparent that, as soon as a crystal-controlled oscillator is used to improve frequency stability, a multiplicity of tubes becomes necessary. At this point the reader is directed to compare the functions of the various tubes, starting with the oscillator, in the modern intermediate-frequency transmitter with the tubes in the modern high-frequency transmitter as labeled in the blocks.

It is evident that, in the high-frequency transmitters [Fig. 104 (d) and Fig. 105 (c) and (d)], frequency doublers follow the buffer amplifier instead of the simple radio-frequency amplifiers used in the intermediate-frequency transmitter [Fig. 104 (c)]. These doublers perform the dual function of radio-frequency amplifier and frequency doubler. The

necessity for each component part of the crystal-controlled transmitter will be explained in the remaining sections of this chapter. Wherever specific mention is made of apparatus construction we are referring to the Western Electric No. 12A transmitter which we consider an example of best design principles as used in intermediate-frequency transmitters. A complete circuit diagram of this transmitter which is of the crystal-controlled type appears in Chap. IX, Broadcast Transmitters.

All of what applies to the intermediate-frequency transmitter also applies to the high-frequency transmitter, the only difference between them being in the L and C values employed and in their propagation characteristics and operation difficulties.

As each component part of the transmitter is studied, its position in the circuit may be seen by referring to the block diagrams before mentioned. For their connection in the circuit as a whole the complete circuit diagrams of crystal-controlled transmitters (Figs. 246 and 253) should be referred to.

104. The Practical Oscillator.—Although the fundamental principles of the oscillator have been previously explained, a more detailed explanation of a practical oscillator, as illustrated in Fig. 109, follows:

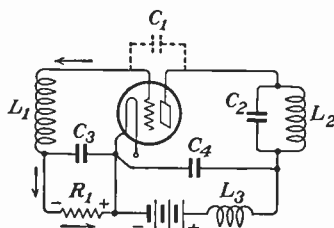


FIG. 109.—Oscillator circuit—series-feed plate supply.

The oscillatory circuit consists of inductance L_2 and capacity C_2 . The plate-filament capacity of the tube will be considered as part of C_2 . The feed-back circuit is the grid-plate capacity C_1 , which is generally quite small, but may be increased, if necessary, by adding an external condenser in parallel with C_1 . However, too large a total value of this capacity is undesirable because the circuit then will be unable to oscillate. In general the output of the vacuum-tube oscillator increases as C_1 increases, up to certain limits depending upon the tube and circuits.

The energy fed back from the plate to the grid circuit through C_1 causes an alternating-current voltage to be built up across inductance L_1 and thus across the grid and filament. The tube then acting as an amplifier develops a correspondingly high voltage in the plate-filament circuit and across the oscillatory circuit giving the latter additional energy and keeping it in sustained oscillation.

The tube obtains its grid bias by means of the voltage drop across R_1 owing to the grid current flowing through R_1 . This current is caused by the positive halves of the alternating-current voltage impressed upon the grid-filament circuit making the grid positive during one half of each cycle. A positive voltage on the grid with respect to the filament causes current to flow from grid to filament through L_1 , R_1 , as indicated by the arrows in Fig. 109. When current flows through R_1 in this direction, the voltage drop across it is such as to make the filament end of the resistance positive and the grid end negative, thus biasing the grid negatively.

On the positive peaks of the alternating-current input voltage the grid is slightly positive because the bias across R_1 does not quite equal these peaks. This condition is necessary in order to have some grid current flowing through R_1 to produce this

negative bias. That is to say, the voltage drop across R_1 is always just sufficient to permit enough grid current to flow in order to produce that grid bias.

Condenser C_3 acts as a by-pass to the input alternating-current voltage and also helps to maintain a steady bias voltage on the grid, for, when the input voltage is negative, no grid current flows and the voltage drop in R_1 is zero. However, C_3 was charged up with the bias voltage during the positive half of the cycle, and it tends to hold that charge and the voltage across it on the negative half. Of course, C_3 discharges through R_1 , but, if each is of large value, such as 0.1 mf. capacity for the condenser and 0.1 megohm for the resistor, the rate of discharge will be slow enough to allow time for the next positive half-cycle to be impressed on the grid.

Condenser C_4 is a by-pass for the plate circuit around the plate supply, and L_3 is a choke coil to prevent alternating current from flowing into the plate-supply circuit. These by-pass condensers and choke coil isolate the input and output circuits of the vacuum tube, thus restricting the coupling between the circuits to that of the coupling capacity C_1 .

Figure 110 illustrates a vacuum-tube oscillator employing inductive coupling for the feed-back circuit. As before, the oscillatory circuit C_2 and L_2 is coupled to the plate-filament circuit of the three-element vacuum tube; R_1 is the grid leak or grid-biasing resistor; and L_4 is a choke coil to prevent alternating current from flowing

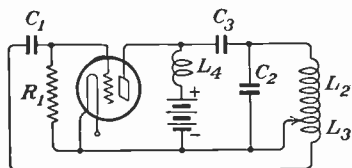


FIG. 110.—Oscillator circuit—parallel-feed plate supply.

in the plate-supply circuit. C_1 and C_3 are blocking condensers to prevent short-circuiting either the bias or plate voltage. L_3 is inductively coupled to L_2 so that alternating current flowing in the latter induces a voltage in the former, thus impressing a feed-back voltage on the grid of the tube through C_1 . The amount of input voltage may be varied to the desired value by adjusting the tap on coil L_2 , L_3 . This will change the resonant frequency of the output circuit, and consequently condenser C_2 must be adjusted to compensate for the change of inductance of L_2 .

One point to be noticed in Figs. 109 and 110 is the difference in the method of connecting the plate-supply voltage. In Fig. 109 the series-feed method is used, while in Fig. 110 the parallel-feed method is employed. Each method has its advantages and special applications depending on circuit and load conditions. In a series feed the plate current flows through the output-circuit inductance. This permits the battery or other power-supply source to be practically at ground potential as far as alternating current is concerned, which is a desirable condition because it reduces losses and prevents intercoupling with other apparatus. However, there is a disadvantage in that the output coil is at a high direct-current voltage above ground, necessitating additional insulation and also extreme care in the method of coupling to an external load.

In the parallel-feed system both difficulties just enumerated are eliminated, but the choke coil is called upon to withstand the full alternating-current output voltage. This means that some alternating current will flow through the choke coil, L_3 , and be lost to the output circuit. This loss may be very great unless the impedance of the coil is high, that is to say, the coil must be wound with many turns of wire to produce a large inductance and at the same time have a low distributed capacity shunting the winding. A properly designed choke coil will offer from 5,000 to 10,000 ohms impedance at the operating frequency, so one or more coils connected in series will prevent practically any current from flowing and thereby greatly reduce the possibility of circuit losses.

There are many possible circuit arrangements for vacuum-tube oscillators, as shown in the preceding chapter, each serving a different purpose under different kinds of operating and load conditions. However, most of them have the same inherent fault, namely, that of frequency instability.

The frequency of the vacuum-tube oscillator is that of the oscillatory circuit, which is dependent upon its own constants, plus the influences of the other circuits of the oscillator and the external load. It has been found that an oscillatory or output circuit with large capacity, low inductance, and very low resistance approaches the ideal circuit for frequency stability. In ordinary oscillators these conditions are desired, but the difficulty of building a coil with extremely low effective resistance, and a condenser of low resistance but of large capacity which must be variable, limits the design and construction, with the result that satisfactory frequency stability using an ordinary oscillator is not commercially practicable for a broadcast transmitter.

A parallel-resonant circuit with the characteristics of high reactance and low resistance is called a *high Q circuit*. Thus,

$$Q = \frac{\omega L}{R} \quad (107)$$

The ratio (Q) of inductive or capacitive reactance to the total resistance of the circuit should be a large value, from 50 to 185, depending on the frequency. In this discussion the total resistance of a circuit means the effective resistance of the coil, condenser, and interconnecting leads, plus the reflected resistance due to an external load, if there is one. Generally, the condensers used in high-frequency circuits have a low effective resistance and consequently this factor is neglected in most calculations. However, this resistance may not be neglected when considering the output circuit of a vacuum-tube oscillator in which extreme frequency stability is desired.

105. The Mechanical Resonator.—While it is difficult to obtain a high- Q electrically resonant circuit for vacuum-tube oscillators, use is made of a peculiar characteristic of certain mechanical resonators to secure the equivalent of such a circuit.

An example of a mechanical resonator or vibrator is a tuning fork. If the fork is given a blow, it will vibrate at a certain resonant frequency and continue to do so until all the energy imparted to it by the blow is dissipated. The frequency at which the fork vibrates is its resonant frequency and is analogous to the resonant frequency of an electrical circuit.

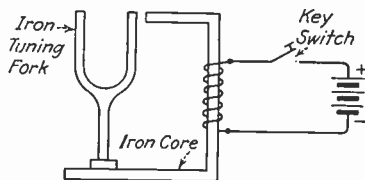


FIG. 111.—Mechanical resonator actuated by external e.m.f.

If the tuning fork is mounted in an arrangement, such as is shown in Fig. 111, it will be excited or operated by magnetic forces instead of the mechanical excitation described above. The coil wound on the iron core has flowing through it, when the key switch is closed, an electric current of sufficient amount to magnetize the core, which attracts one prong of the tuning fork to it. If the key is opened, current ceases to flow, neglecting the momentary surge and arcing at the switch points due to collapsing of the field, and the core becomes demagnetized and releases the tuning fork. If the key is closed and opened rapidly, the fork will be attracted and released accordingly.

If we assume that the key is being closed and opened more and more rapidly, the fork responding to each operation, the number of vibrations of the fork will increase until finally this number will equal the resonant frequency as determined by mechanical excitation. If the current is stopped entirely, the fork will continue to vibrate for a time until its energy is dissipated.

The important point in this phenomenon is that below the resonant frequency considerable magnetic pull is required to operate the fork and, as soon as the current ceases, the fork will stop vibrating, while at the resonant frequency very little magnetic pull is required and, if the current is stopped, the fork will continue to vibrate for a time at its own frequency. This apparatus arrangement, operating at the resonant frequency of the tuning fork, is another form of mechanical resonator. The next step will be to consider the process reversed.

In Fig. 111, let us assume the iron core to be permanently magnetized and the battery and key to be replaced by a sensitive meter such as a zero-center galvanometer or an oscillograph element in series with the coil, as in Fig. 112. Now, if the prong of the tuning fork is suddenly moved toward the upper end of the iron core, there will be a deflection of the meter in a given direction, and, if the fork is suddenly released, it will spring back to the original position and the meter will deflect in the other direction. The reason for this action is that, as the fork moves toward the iron core, the reluctance of the magnetic circuit was changed which caused a change in the magnetic flux traversing the magnetic circuit constituted by the core. The coil being wound on the core, and thus linking the flux, has a voltage induced in it, owing

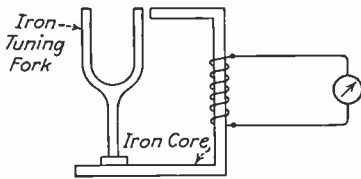


FIG. 112.—Mechanical resonator as a producer of e.m.f.

to the change in flux and current flow in the electric circuit, which causes the meter to deflect. The other half of the action is similar but in the opposite direction. The fork moving away from the core increases the magnetic reluctance, decreasing the flux, and this decrease again induces a voltage in the coil but in the reverse direction with resultant reversal of the meter deflection. As soon as the fork has come to rest and there is no longer a change in the

magnetic circuit, the meter will again return to zero reading.

Now, if the fork is moved toward the core and released more rapidly, the above action is repeated and the meter is seen to oscillate about the zero point at a frequency or frequency period corresponding to that of the movement of the fork. Consider the mechanical excitation of the fork increasing more and more until finally the number of vibrations equals its resonant frequency, when it will vibrate easily with little excitation and will continue for a time with the excitation entirely removed.

The purpose of this experiment is to demonstrate that the vibrating tuning fork indirectly causes the production of alternating currents in the electric circuit, and therefore the whole process is reversible. That is, with the first arrangement in Fig. 111 the closing and opening of the key started and stopped the flow of current in the coil which through magnetic action set the fork into oscillation. And, conversely, the vibrating tuning fork through magnetic action sets up alternating currents in the electric circuit as indicated by the meter in the arrangement of Fig. 112.

106. The Crystal Resonator.—Still another form of mechanical resonator is the piezo-electric crystal. Several crystalline materials, such as quartz, tourmaline, Rochelle salt, and cane sugar, have the

property of producing an e.m.f. when the crystal is mechanically strained (compressed or expanded). Furthermore, this action is reversible, so that, when an electric charge is impressed upon the crystal, the latter changes its shape. This remarkable characteristic is called the *piezo-electric effect* and is present in all the so-called piezo-electric materials to a greater or less degree, dependent upon the material, upon the form in which it is cut, and the manner in which the electric field is impressed. Rochelle salt is the most active in this property, but quartz has more mechanical advantages in that it is practically unaffected by light shocks, aging, and moisture. There is another advantage which is even more important; it is, that, due to its extreme hardness and low internal friction when vibrating, it has low damping and consequently a high-Q characteristic. This property, as we have seen before, is the chief characteristic necessary in the oscillatory circuit of a vacuum-tube oscillator where great frequency stability is required, and it is here that the quartz crystal plays the major role.

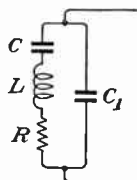


FIG. 113.—
Electrical
equivalent
circuit of piezo-
electric crystal
resonator.

The piezo-electric crystal as a resonator may be considered equivalent to an electrical circuit such as is shown in Fig. 113. The series condenser C represents the rigidity of the crystal, the inductance L its mass, and R the internal resistance due to internal friction. The parallel capacity C_1 is mostly due to the capacity effect of the metal electrodes of the crystal holder acting like a condenser with the crystal as the dielectric. In practice, added to this is the capacity of the external connections and the grid-filament capacity of the vacuum tube when used.

Before proceeding further, one point must be clearly understood. Although the piece of quartz used as a resonator is generally called a "crystal," still it is not a whole crystal but only a section of one cut to certain dimensions and specifications. This section may be a thin slab in rectangular form, or it may be cut like a disk. The smaller, thinner sections used for very high frequencies are circular cut similar to a coin. Still another form, which is used for precision work, but not in commercial transmitters, is a flat ring. So let us proceed from here by calling the piezo-electric crystal a *quartz plate*, meaning a thin slab of rectangular or circular form, cut from a high-grade quartz crystal along a particular axis to make it more active piezo-electrically.

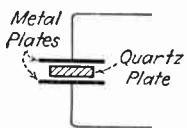


FIG. 114.—
Method of making
connections to
quartz plate.

Figure 114 shows schematically a method of making electrical connections to the quartz plate. It is usually placed in a horizontal position between two metal electrodes or plates. The lower electrode furnishes a base for the quartz plate and is an integral part of the whole mounting. The upper electrode may be spaced above the plate, touch it, or be

clamped to it. The crystal, like the tuning fork, must be permitted a certain amount of mechanical freedom in order to oscillate. The early practice was to permit the upper electrode to rest lightly upon the surface of the plate, a flexible connection being made to the electrode. The difficulty with this method was that the electrode would move about, and small particles of foreign matter would settle under it resulting in a change of spacing between the plate and the electrode. This condition caused considerable variation in frequency. To avoid such variations the quartz-plate holder is now constructed so as to clamp the plate rigidly in one position. Each electrode is made with an uneven surface of rises and depressions. The rises make contact and clamp the plate, while the depressions provide air spaces between plate and electrode, which allows the plate to oscillate or vibrate.

Figure 115 shows a schematic diagram of a vacuum-tube oscillator incorporating a quartz-plate resonator. As previously explained, the tube with its associated circuits, such as filament and plate supply and automatic bias by means of R_1 , and tuned output circuit, constitutes an amplifier. Feed back to the grid circuit is accomplished by the

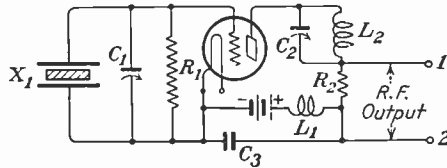


FIG. 115.—Oscillator with quartz-plate resonator.

grid-plate capacity of the tube. The resonant circuit that determines the operating frequency of the oscillator is the quartz plate X_1 which is connected across the grid and filament of the tube. The condenser C_1 is of small capacity and is shunted across the quartz plate to obtain a vernier adjustment of the operating frequency by changing the resonant frequency of the equivalent electrical circuit of the resonator. The resistor R_2 acts as a load to the amplifier tending to stabilize it against external load changes and also to provide a means of coupling for the load. This resistance is only a few hundred ohms so that variations in coupling from a high impedance load do not affect the oscillator appreciably.

When the plate voltage is suddenly applied to the vacuum tube, a surge is produced in the output circuit charging C_2 , which tends to discharge through L_2 similar to the operation of an ordinary oscillatory circuit. This action in the output circuit causes a voltage to be built up across the plate and filament of the tube which, through the feed-back coupling of the grid-plate capacity, impresses a voltage upon the grid-filament circuit and across the quartz plate. The voltage in the latter circuit sets up an electrostatic field around the plate causing it to change shape slightly. As soon as the impressed voltage dies down, the quartz plate returns to its normal shape and in so doing produces a voltage across the electrodes and across the grid and filament of the vacuum tube. This voltage is amplified in the plate circuit, and some of the energy is fed back again to the grid circuit and the quartz plate which causes the latter to vibrate or oscillate, thereby applying a varying voltage to the grid. This action results in a continuous building-up process of the oscillations until the vacuum-tube oscillator is delivering its maximum power to its own and the external load for a given set of

operating voltages and circuit constants. Actually the output circuit is not tuned to the exact frequency but to a slightly higher one so that it is somewhat inductive at the operating frequency. This condition provides greater stability and also aids the quartz plate to start vibrating more readily.

The main factors which determine the operating frequency of the vacuum-tube oscillator and also its stability are: the type of cut and mode of vibration, whether longitudinal or lateral, of the quartz plate; its thickness; the relationship between the plate and the electrodes; the amount of capacity externally connected across the electrodes; and the temperature of the quartz plate. Other factors, such as power supply and external load conditions affect the frequency only slightly in that they cause minor variations of the load across the quartz plate itself.

The type of cut and mode of vibration of the plate depend upon the frequency desired, the type of operation and the design of the oscillator. These are fixed properties in any particular plate. The thickness of the plate required depends principally upon the frequency of operation of the crystal. The relationship between plate and electrodes has become a fixed quantity owing to the clamping process explained above.

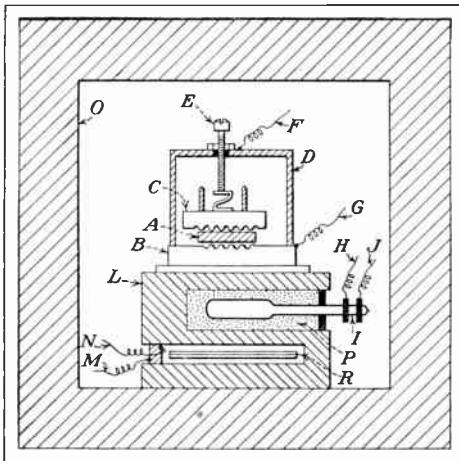


FIG. 116.—Temperature chamber for quartz-plate resonator (see text, Sec. 107, for an explanation of parts).

The external capacity, once the oscillator is designed, built, and adjusted, is a question of the grid-filament capacity present in the tube and the vernier condenser C_1 . The former is generally of the order of 5 or 10 mmf. and changes very little as a result of fairly uniform construction in manufacture. The vernier condenser is designed to be adjusted for small variations of frequency of the order of 10 or 20 cycles, but once adjusted it is seldom varied. This leaves the temperature as the one remaining factor, and it is actually of major importance.

The temperature of the quartz plate is maintained at a nearly constant value by placing the plate and its holder, together with a heater and thermostat, in a heat-insulated chamber. The chamber may be a metal box carefully lined inside with an insulating material such as Celotex. It is sometimes advisable to enclose this entire unit along with the vacuum tube and associated circuits in another insulated chamber, but this is only for very accurate standards.

107. The Temperature-control Chamber.—Figure 116 shows an arrangement of a temperature chamber for a quartz-plate resonator.

This is the Western Electric type which is used in many broadcast stations. The quartz plate *A* rests upon the lower electrode *B* and is held in place by the clamping action of the upper electrode *C* with its spring and adjusting screw *E*. A locknut fixes the screw once it is adjusted. The shell *D*, which surrounds this assembly and fastens to the lower electrode, is an insulating material such as Isolantite. The lower electrode is connected thermally to the heater unit through a thermal filter. The heater unit *L* contains the thermostat *I* and heater element *R*. Graphite powder *P* holds the thermostat in place, and leads *H* and *J* make electrical connection to the mercury contacts within the

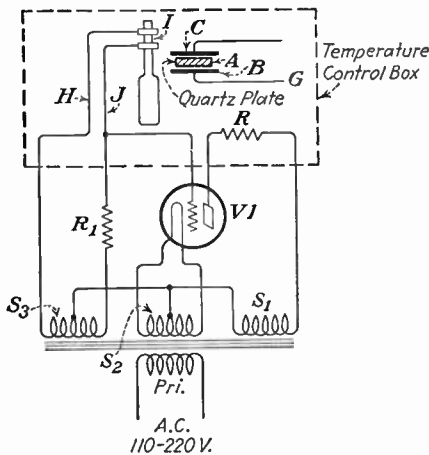


FIG. 117.—Schematic diagram of connections in temperature chamber.

the vacuum-tube oscillator and also to the temperature-control circuits, shown schematically in Fig. 117.

The purpose of the control circuit is to provide heat for the quartz plate and maintain the temperature of the plate as nearly constant as possible. Heat is not essential to the operation of the quartz resonator, but, as room temperatures may vary as much as 50°F., the simplest method is to heat the plate considerably above room temperature (to about 90°F.), so that these changes are less effective, and then maintain a constant value. As shown in the schematic drawing, heat is supplied by passing current through the resistor *R*. This current is obtained from the transformer *S*₁, the circuit being completed through the plate filament of the control vacuum tube *V*₁. This tube is generally a gaseous-type rectifier with a low internal voltage drop and resembles the ordinary high-vacuum three-element tube in that a control grid is used.

The control action is not one of regulating heat to meet varying conditions but is an intermittent process. Normally the circuit through the thermostat is open, because the mercury column within is normally low. This condition leaves the grid of the vacuum tube connected through resistor *R*₁ to what we shall call the positive side of the transformer *S*₃ with respect to the mid-tap which is connected to the filament through transformer *S*₂. By "positive side of the transformer" is meant that

The heater element *R* may be of the wire-wound or grid type, carefully insulated electrically from the copper casting of the heater unit, with connecting leads *M* and *N* brought outside. A thermal-insulating housing *O* completely encloses all this apparatus tending to maintain the internal temperature constant. The whole unit is assembled in a metal box designed to be mounted in the radio transmitter. All the leads including those from the electrodes are brought through the box to a terminal strip where connections are made to the grid and filament of

the windings are so phased that during the half of the alternating-current cycle, when the plate of the tube is made positive, the grid is also made positive, all with respect to the filament. Under this condition the tube is conducting, and current flows around in the plate circuit, heating the resistor R . During the other half of the cycle, both plate and grid are negative and no current flows. Thus we have a half-wave rectifier supplying current to the heater, the tube being conducting on the positive half-cycle because the control grid is also positive.

After the heater element has been in operation for some time, the temperature inside the chamber rises. Then the mercury in the thermostat expands, short-circuiting the contacts in the stem. This completes the circuit of the entire winding of transformer S_3 through resistor R_1 and the thermostat contacts. All the voltage drop is consequently across R_1 , the control grid is now connected to the left or "negative" side of the transformer, that is, the grid is negative whenever the plate of the tube is positive, it is biased to plate-current cut-off, and, of course, no plate current flows. On the negative half of the cycle the grid becomes positive, but still no plate current flows because the plate of the tube is now negative. Therefore, the action

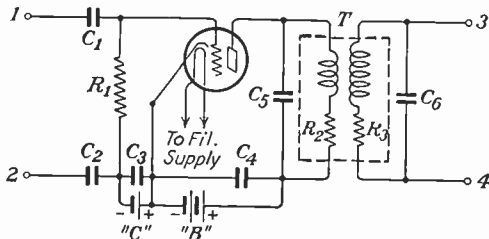


FIG. 118.—Circuit of buffer or first amplifier.

of the thermostat has been to change the phase relation of the voltage on the control grid and thus indirectly prevent further heating of R .

Eventually the heat in the chamber will escape through the walls, and the temperature inside will drop. The mercury in the thermostat also drops, opening the contacts in the stem and returning the control tube to its original condition, and the heating of R begins again. Thus the whole cycle consists of heating R and then permitting it to cool. The heating period may take 30 sec. and the cooling period 50 sec., depending upon many factors of design of the whole system. One may think the quartz plate would also be alternately heated and cooled, but this condition is largely prevented by a specially designed heat filter which tends to iron out these changes and provide a constant flow of heat to the plate.

108. The Buffer Amplifier.—The main purpose of this stage, shown in Fig. 118, is to function as a first voltage amplifier and at the same time isolate the oscillator from the succeeding amplifiers in the radio transmitter as far as the carrier frequency is concerned. The buffer amplifier usually consists of a single vacuum tube biased to at least the I_p cut-off point and so operated that the positive peaks of the radio-frequency voltage applied to the grid and filament of the tube from the oscillator do not exceed the grid-bias voltage. Under this condition the amplifier draws no grid current and consequently there is no changing load on the oscillator. If, however, an appreciable amount of grid current is drawn with the resultant variations of load on the oscillator tube,

there would be variations in the frequency owing to the slight changes of the resonant frequency of the quartz-plate resonator. Hence, by employing a buffer-amplifier stage, the oscillator is permitted to work into its own fixed-load resistance, supplying only a radio-frequency voltage to the buffer amplifier.

In the schematic diagram of the vacuum-tube oscillator (Fig. 115) is seen the load resistance R_2 across which is developed the carrier-frequency voltage which is to be applied to the buffer amplifier. Figure 118 is a schematic diagram of this amplifier. The input voltage is applied to terminals 1 and 2, and is impressed on the grid and filament through the blocking condensers C_1 and C_2 and the by-pass condenser C_3 which shunts the grid-biasing potential. The blocking condensers C_1 and C_2 prevent a short circuit of the grid-biasing resistor R_1 by the load resistance in the oscillator. The negative bias is applied to the grid through resistance R_1 which is of a large value, such as 0.5 megohm, to prevent excessive losses in the input radio-frequency current. Owing to the fact that no grid current is permitted to flow in the grid circuit, it is advisable to use a large resistance for R_1 so that the input impedance is kept as high as possible. Condenser C_4 is a by-pass for the plate-supply battery.

The output or load circuit of the buffer amplifier may be of the type illustrated in Fig. 118 which consists of an air-core, shielded, fixed-tuned transformer T for coupling to the succeeding stage, at the terminals 3 and 4. This type of transformer offers a fairly constant impedance into which the tube may work and has a good transmission characteristic or frequency response over a considerable band width. With proper design, the transmission characteristic may be constant to within 25 per cent over the entire broadcast band. It is generally unnecessary to neutralize this amplifier because of its relatively low power and gain. The primary of the transformer is tuned to a given frequency by the condenser C_5 , while the secondary is tuned to a slightly different frequency by the condenser C_6 which consists of a fixed condenser and the grid-filament capacity of the succeeding tube. The result of the two tuned circuits when coupled together is a double-resonant frequency looking into the primary circuit from the tube. The peaks of the double resonance may then be caused to spread apart by the proper adjustment of the coupling between the two windings and the peaks then flattened out by the resistances R_2 and R_3 of the windings. Hence the resultant frequency-response curve results from two widely separated very flat resonance curves which have merged into each other effectively, thus producing a uniform frequency characteristic over a wide band. Once this adjustment has been made, the transformer remains unchanged and the condensers C_5 and C_6 may be fixed capacities.

109. The Second Amplifier.—The general block diagram of a radio transmitter as illustrated in Fig. 104(c) indicates a second amplifier

(schematic diagram shown in Fig. 119) coupled to the buffer stage. It was shown that the buffer amplifier served a double purpose; first, that it is used chiefly to isolate the oscillator from the other circuits in order to obtain a maximum frequency stability, and, second, to function as a low-gain voltage amplifier. The second amplifier, however, must operate not only as a voltage amplifier but also as a power amplifier as shown in Fig. 119.

A true voltage amplifier draws no power from the preceding stage and delivers practically no radio-frequency power to the succeeding stage. There is, of course, some quantity of power loss due to the coil and the circuit resistance, but these are reduced to as low a value as possible in the design and construction. The voltage amplifier is usually a high-gain device, in which a small input voltage is received and a large output voltage is delivered to the succeeding stage. The coupling transformer is generally tuned in the primary and sometimes also in the secondary, and the primary-to-secondary turn ratio is quite large.

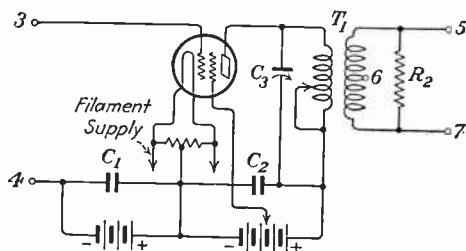


FIG. 119.—Circuit of the second amplifier.

A true power amplifier requires a large value of input voltage and considerable input power while delivering a large value of output power with some voltage gain. The output transformer may therefore have a relatively few number of turns on the secondary and only enough primary inductance to obtain the proper impedance match for the tube. Both of the windings are generally tuned where high-power amplification and a maximum transfer of energy are required. The output transformer for this type of amplifier, therefore, may be a step-down transformer with few turns but using large wire.

It has previously been shown that voltage amplifiers employ vacuum tubes with a high amplification factor (μ) and a large plate resistance (R_p), while at the same time the impedance of the output circuit is made as high as practical design will permit. Power amplifiers, on the other hand, generally employ vacuum tubes with a low amplification factor and plate impedance in order that they may be capable of handling large amounts of power with minimum distortion. The output impedance is usually adjusted to match the plate resistance of the tube, in order to obtain the maximum transfer of energy from the tube to the output.

For certain purposes such as in a modulated amplifier, the above conditions may not be adhered to; that is, this stage may be both a voltage and a power amplifier, requiring a small power input and delivering a fairly large power and voltage output, but all with the chief characteristic of linear amplification in which the amplified output is a faithful reproduction of the input voltage. This type of amplifier will not be so efficient as a true voltage or power amplifier, and its overall gain will be lower than if it were so used.

In the radio transmitter now under consideration, the second amplifier falls in the third class; that is, it is used for both voltage and power amplification. For reasons yet to be explained, it is not used for the purpose of providing a linear amplifier.

In the schematic diagram of this amplifier as given in Fig. 119, it is seen that a screen-grid vacuum tube is employed. The radio-frequency input voltage is obtained by connecting terminals 3 and 4 to the corresponding terminals on the buffer amplifier. The screen-grid tube is used for two reasons: one is that it is unnecessary to neutralize this stage when this type of tube is used, and the other is that a means is then provided for controlling the input to the modulating stage to follow, and also for controlling the radiated output of the transmitter.

When it is desired to increase the radio-frequency output of the second amplifier, the screen-grid voltage is increased slightly, effectively reducing the plate resistance and increasing the plate current and output of the tube. Conversely, when the output is to be decreased the screen voltage is lowered.

Owing to the fact that the modulating stage follows the second amplifier, it is necessary to dissipate a certain amount of the radio-frequency energy in the load resistance R_2 in order to obtain the proper voltage across terminals 5 and 7 of the transformer T_1 . Hence the second amplifier must deliver some power to the load resistance (and perhaps some to the input of the modulating stage depending on the type of modulation) and also amplify the voltage impressed across its grid and filament. Screen-grid tubes owing to their design have a fairly high plate resistance and amplification factor making them ideal voltage amplifiers. When this type of a tube is called upon to deliver power to a load, it needs a larger input voltage and some power, because of the large values of grid current resulting from the high input voltage.

Figure 119 shows the usual by-pass condensers C_1 and C_2 , plate supply, and output transformer T_1 . The primary of T_1 is tapped in order to cover the required frequency band in tuning in conjunction with the variable condenser C_3 . The load resistance R_2 is a specially wound unit designed to have practically no inductance or capacity. The value of R_2 is determined by the voltage to be impressed on the grid and filament of the succeeding stage and by the power to be dissipated in the resistance from the second amplifier. Hence, assuming that the succeeding stage required an input voltage of 100 volts, and assuming that the second amplifier was to deliver 10 watts to the resistance, then, by Ohm's law ($E = IR$), and the power equation ($W = EI$), the resultant equation ($R = \frac{E^2}{W}$) gives $R = \frac{100^2}{10} = 1,000$ ohms. Therefore, when the second amplifier delivers 10 watts to the resistance R_2 , there will be built up across it 100 volts which is impressed on the grid and filament of the modulating stage.

110. Harmonic Producers: Frequency Multipliers.¹—Systems which are rich in the generation of harmonic frequencies may be used to advan-

¹ For the sake of continuity, if only broadcast transmitters are being studied, it is desirable to leave this section for later and separate consideration as Sec. 111 appropriately follows Sec. 109 at this point.

tage in radio-frequency amplifiers for the purpose of *frequency multiplying*.

It has been shown that all single-tube amplifiers possess this characteristic due to the non-linearity existing in the plate-current curve when signal e.m.fs. are applied to the grid. Although these conditions are highly undesirable in audio-frequency amplifiers, they may be used to advantage where frequency doubling or tripling is required. In reality, therefore, any amplifier which is adjusted to the point at which high plate-distortion results may be used as a frequency multiplying device.

There are two distinct types of frequency multipliers, namely, that in which the distortion takes place in the plate circuit, and the other in which the distortion takes place in the grid circuit. The former is known as a plate-distortion harmonic generator and the latter as a grid-distortion harmonic generator.

In the plate-distortion method the plate potential should be considerably higher than for normal operating conditions, and the grid bias should be so adjusted that the tube is operating slightly in excess of the I_p-E_c cut-off value so that the input signal voltage will not cause any grid current to flow. The grid-bias voltage, incidentally, is one of the most important adjustments for proper harmonic generation and should be carefully varied for each harmonic change required. The conditions for harmonic generation are therefore almost identical with those of the adjustment of a class B amplifier with the exception that the higher plate-voltage and grid-biasing conditions produce a much more asymmetrical wave form. This is due to the fact that, if the input signal voltage is brought up to the point at which it equals the grid-bias voltage at peaks, it will produce short-duration pulses in the plate circuit.

In the grid-distortion method a series impedance is connected in the grid circuit to form a parallel-resonant circuit properly adjusted to the harmonic frequency that is to be generated. Now, since a parallel-resonant circuit will represent a high impedance in the grid circuit to the resonant-harmonic frequency, the exciting grid voltage will develop a considerable potential difference across it. Now, if the grid bias is adjusted to the point in which a very small grid current will flow during the positive half of the cycle, this small current will be of a highly damped character due to the dissipation of the high series impedance. Consequently, the resultant grid signal voltages developed across the tube will be highly damped. Thus, if the input grid voltages represent short-duration pulses (highly damped pulses are rich in harmonic content), these will be amplified in the plate circuit in accordance with the general considerations of vacuum-tube circuit amplifying ability.

Perhaps the most frequent application of the frequency-multiplying arrangements is that of the quartz-plate oscillator for high-frequency transmission, although the principle may be applied to any conventional type of oscillator-amplifier system.

For example, if it is desired to operate a transmitter at a frequency of 15 megacycles (20 meters), the oscillator must be capable of delivering sufficient power to excite the grids of the succeeding radio-frequency amplifying stages. This would require an oscillator capable of developing at least 5 watts of output power. If, therefore, an oscillator tube is to generate this amount of power, a considerable voltage must be developed across its input circuit through feed back. Consequently, if a quartz crystal is used to control the frequency of the oscillations to be generated, the strain developed across it may subject it to rupture and possible splitting or cracking. It will be remembered, that a crystal, to operate at very high frequencies, must be very thin in its structure. It is, therefore, possible to use high-frequency crystals with assurance of stability, only if the power output requirement is very small. It is necessary, therefore, when it is desired to use a crystal-controlled oscillator for

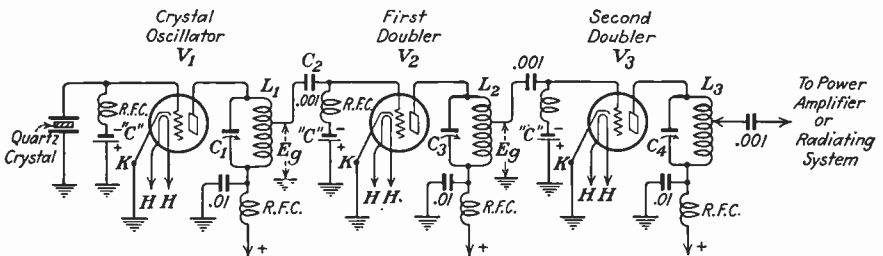


FIG. 120.—Frequency-doubling amplifier circuit.

frequencies of 15 megacycles or more, that a system of frequency doubling be used which will enable the crystal circuit itself to operate at a lower frequency and thereby permit the use of a thicker crystal.

Figure 120 illustrates a typical circuit arrangement of a crystal-controlled oscillator and frequency-doubling amplifier.

Let us assume the crystal to be properly ground to oscillate at a frequency of 3,750 kilocycles (80 meters), and to be connected in the oscillator circuit with tube V_1 . The plate circuit of the oscillator L_1C_1 is adjusted to resonance with the crystal frequency (80 meters). The voltage developed across L_1C_1 may then be applied to the input of the first frequency-doubling tube V_2 through the coupling condenser C_2 . The amount of grid-excitation voltage being determined by the position of the tap on L_1 , $E_g = IZ$. Now, if the plate circuit L_2C_3 of V_2 is adjusted to the second-harmonic frequency or 7,500 kilocycles (40 meters), and the plate and grid-bias voltages adjusted in accordance with the grid- or plate-distortion methods described, the fundamental frequency voltage E_g of V_1 will develop in the plate circuit L_2C_3 an oscillating current of 7,500 kilocycles (40 meters). If the voltage developed across L_2C_3 is insufficient to excite the grid of V_3 , a buffer stage is usually used after

the oscillator, which delivers a larger power at the fundamental frequency. A large size V_2 may then be used which can properly excite V_3 . The voltage now developed across L_2C_3 at the frequency of 7,500 kilocycles (40 meters) will, as before, develop a frequency at the second harmonic of 7,500 kilocycles or 15 megacycles (20 meters), in L_3C_4 , provided, of course, that the plate and grid-bias voltages are applied in accordance with harmonic-generation procedure, and that the load circuit of V_3 L_3C_4 is adjusted to 15 megacycles. Now, since this is the desired frequency to be transmitted, the output of L_3C_4 may be coupled to the radiating system or it may be amplified through a number of voltage and power-amplifier stages to obtain the desired antenna output.

Great care must be taken that all stages are well shielded and neutralized, and all circuits properly isolated by grid- and plate-filtering systems. Doublers do not generally require neutralizing.

If the output circuit of a frequency-multiplying tube is to provide other than a single-harmonic frequency, the output load is generally an untuned inductance so that any desired harmonic may be obtained. The impedance of this inductance, however, should be made variable by taps so that the load may be regulated to the proper impedance match for the particular harmonic desired.

111. Modulation.—With the exception of the vacuum-tube oscillator with its crystal control for high-frequency stability, the modulated-amplifier stage is probably the most difficult part of the transmitter to design, adjust, and operate from an electrical standpoint. It is here that the single radio-frequency wave must be converted into a very complex wave by the incoming speech or music from the audio amplifiers in such a manner that when the signal is demodulated by the receiver, the resulting speech or music is a faithful reproduction of the original program or other speech.

By modulation is meant the process whereby the frequency or amplitude of the radio wave is varied in accordance with an audio wave. There are two ways of modulating a radio wave, by either frequency or amplitude modulation, but, commercially, only the latter is employed in modern radio telephony. Frequency modulation is that method in which the fundamental carrier frequency is made to change in accordance with an audio frequency, for example, by causing an audio frequency to vary the capacity of the tuned circuit of the vacuum-tube oscillator, thereby varying the carrier or fundamental frequency. This method is employed to secure interrupted continuous waves (*ICW*) on marine transmitters, and will be explained in Sec. 252.

In amplitude modulation the carrier frequency remains constant, but its amplitude is made to change in accordance with the audio frequency. The carrier wave normally has a certain value or amplitude. Thus, when the audio-frequency wave is impressed on the modulating

stage, a sort of valve action occurs and the carrier wave becomes larger or smaller in accordance with the action of the audio wave. For full modulation, when the positive peak of the audio wave reaches its maximum, the peaks of the carrier wave become double the value reached when no modulation is occurring; and when the other half of the audio wave reaches a maximum, that is, in the negative direction, the carrier-wave peaks are reduced to zero.

Figure 121 shows the conventional method of illustrating graphically what occurs during modulation. The upper wave represents the alternations of the carrier-frequency current while the lower wave represents the alternations of a single audio-frequency current. Along the horizontal time axis the distance *AB* indicates the condition existing before the audio wave is applied. The carrier wave is then seen to alternate about the zero axis with peak values of current remaining constant at an amplitude equal to *P*. At the point *B* on the time axis, the audio wave starts to

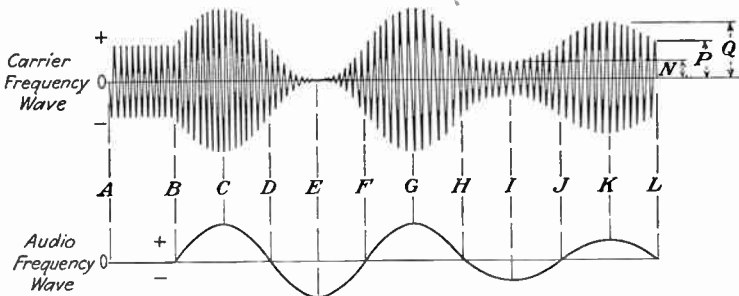


FIG. 121.—Modulating effect of audio-frequency wave on carrier frequency.

increase in the positive direction causing an increase of the carrier-frequency output of the modulating stage. At point *C* the peak of the audio wave causes the maximum output and then starts decreasing to *D* where the carrier is again normal. The audio wave now starts in the negative direction, decreasing the carrier still further, until point *E* is reached where the carrier output is zero. Further on, at the point *F*, the audio wave is zero, having completed one alternating cycle, and the carrier is again back to normal and the entire operation is ready to be repeated.

112. Modulation Percentage.—The audio wave was chosen of such amplitude during the time interval *B* to *H*, that full, or 100 per cent, modulation resulted. Now if the amplitude of the audio wave were decreased to half that value, such as is shown during the time interval *H* to *L*, then the amount of modulation would be only half of that first obtained. The measure of the amount the carrier wave is changed is expressed by a value *M* called the *percentage of modulation*. This is a measure of the depth of the envelope of the modulated carrier and is expressed by the equation:

$$M = \left(\frac{I_{max} - I_{min}}{2I_0} \times 100 \right) \text{ per cent} \tag{108}$$

where I_{max} is the maximum peak value of carrier current (indicated by the special commonly used symbol Q in Fig. 121); I_{min} is the lowest peak value of current (indicated by N); and I_0 is the normal peak value of current when no modulation is occurring (indicated by P).

Taking the time interval B to F on the modulated carrier wave, the amplitude Q is seen to be twice the normal value P , while the amplitude N is considered to be zero. Substituting these values in the formula above, the percentage of modulation is found by the equation:

$$M = \frac{Q - N}{2P} \times 100 = \frac{2P - 0}{2P} \times 100 = 100 \text{ per cent} \quad (109)$$

If the audio wave has a smaller amplitude than that given above, say, for example, the amplitude in the interval H to L , then the percentage of modulation will be less, that is, 50 per cent. Therefore, the percentage of modulation may be any value from 0 to 100 depending upon the amount of audio frequency applied to the modulating stage, but it can never exceed 100 without introducing distortion. If more audio frequency is applied to the modulating stage than the amount necessary to produce 100 per cent modulation, then the carrier is said to be overmodulated and the amount of distortion caused in the audio-frequency component is increased as the overmodulation is increased.

113. Carrier and Side-band Relations.—The theory of modulation may best be understood by considering two important viewpoints, namely, one in which the modulation process may be visualized as a wave of constant frequency with varying amplitude (amplitude modulation), and one in which the wave is expanded into components of the carrier and the side-band frequencies.

Amplitude modulation as shown in Fig. 121 is unquestionably a very definite picture, and serves to illustrate how a wave form may be varied in its amplitude by variable resistance or reactance characteristics. This condition may be readily understood by considering the explanations in Sec. 111 and 116. It is simply a method of expressing the physical change of the wave shape into various amplitudes, and is brought about by the application of an audio-frequency wave upon the carrier frequency. Regardless of whether or not this amplitude change is brought about by grid or plate modulation, it is still visualized as amplitude modulation as shown in Fig. 121.

The side-band viewpoint, on the other hand, may in some cases present a more desirable picture since it illustrates that during the process of modulation, the modulated signal appears as three components (or more) on three (or more) different frequencies. This viewpoint may be clearly seen from the illustrations in Fig. 122 (a) and (b).

Let us assume a carrier frequency of 1,000 kc./s. modulated by a steady audio frequency of 7,500 cycles. This is generally the highest audio frequency transmitted by broadcast stations, except in special cases where high-quality telephone lines of 8,000 cycles or better are used. It will be seen from both illustrations that the impressed audio frequency will result in the generation of two additional frequencies, one of 1,007.5 kc./s. and another of 992.5 kc./s. These are known as the *upper* and *lower side bands*, respectively. Now, since this interaction of frequencies occurs in the transmitting circuit, these reactions will produce a resultant amplitude change in

the output wave form and its form will be as illustrated in (e) of Fig. 122 (a). In other words, the resultant wave shape (e) represents the instantaneous sum of the three radio frequencies (b), (c), and (d). It is this wave which represents the true picture of the modulated output emanating from the transmitter during the process of modulation. The ratio of the sum of the two side bands divided by the carrier frequency is generally referred to as the *modulation factor*, expressed

$$m = \frac{I_1 + I_2}{I_c} = \frac{2I_s}{I_c} \tag{109A}$$

Definite proof that side bands exist can be obtained by placing a sharply tuned calibrated radio-frequency amplifier with a crystal filter and an output indicator of the oscillograph variety in the vicinity of the transmitter. If the receiver is tuned to the unmodulated carrier frequency, and the received frequency measured and noted, and the carrier then modulated at 7,500 cycles, and the circuit carefully tuned to a point 7,500 cycles above or below the carrier frequency, a resonant peak at the two side bands will be indicated. Hence, if the energy for each of the three frequencies is plotted, a characteristic such as is illustrated in Fig. 122 (b) will be obtained.

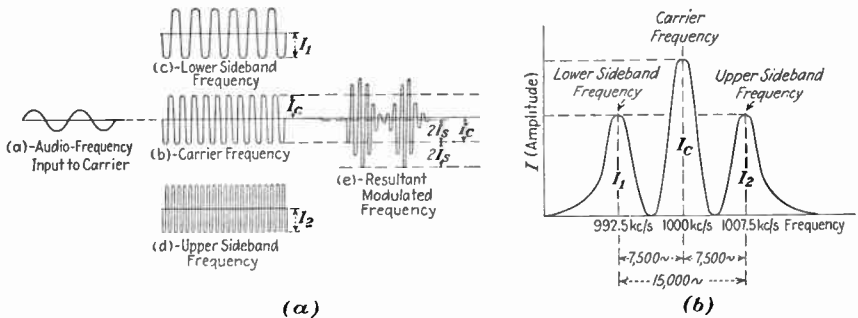


Fig. 122.—Relationship of side bands to carrier frequency.

It is the interaction of the side-band frequencies with the carrier frequency which produces the resultant or beat frequency. Hence, in the case of 7,500-cycle modulation there are side-band frequencies of 7,500 cycles above and 7,500 cycles below that of the carrier frequency or a band width of 15,000 cycles, but the actual resultant frequency due to the interactions is 7,500 cycles. The intelligibility, therefore, is entirely due to the side-band frequencies and their reactions with the carrier and not to the carrier frequency alone.

The power contained in the side bands depends upon the variations in the amplitude of the wave. Consequently if the modulation cycle is complete, or 100 per cent, the maximum amount of power will be contained in the side band.

114. Modulating Methods.—Modulation as it is applied to broadcast transmitters is always accomplished by means of one or more vacuum tubes. In general, one or more of these tubes are employed in the modulating stage or modulating amplifier (which is a linear amplifier of the carrier frequency, and the one in which the modulation actually occurs). The modulator itself is really the last stage in the audio-

amplifier system and supplies the audio-frequency voltages to the modulating amplifier.

There are two general classes of modulation each of which may be subdivided into distinct groups.

The first class will be called *grid modulation* because the audio frequency is applied to the grid of the modulating amplifier. In this class the actual modulation may occur in the grid circuit of the vacuum tube in which case it is called *grid-current* modulation, or it may occur in the plate circuit, the tube acting as a double amplifier, when it is called *grid-bias* modulation.

The second class will be called *plate modulation* because the audio frequency is applied directly to the plate of the modulating amplifier and practically all the modulation occurs in that circuit. In this class there are a number of circuit arrangements possible, the principal ones being the *Heising system* and the *modified Heising system*. In the former, the plate supply to both the modulated amplifier and the modulator is obtained from a common source and is fed to them through a high-inductance, low-resistance choke coil, while in the latter system each tube has a separate plate supply and the coupling between the two plate circuits is obtained by means of an audio transformer or large condenser.

Another method of classifying modulation is to consider it to be either of the high- or low-level method. *Low-level modulation* means that it occurs in one of the low-power stages and that all succeeding stages are linear amplifiers of the modulated carrier. This method may employ either grid or plate modulation as explained above. The main advantage of this type is that the audio-frequency amplifier in the transmitter may consist of only a few stages of low-level amplification; that is, very little audio power will be required.

High-level modulation means that the plate circuit of the last radio stage is modulated. This definition automatically establishes plate modulation as the class employed. The advantage of this type is that all the preceding radio stages are pure radio-frequency amplifiers of the carrier wave, are not necessarily linear, and may have a high gain in each stage. Grid-bias modulation when employed in the last radio stage might be considered a modified form of high-level modulation, but when used in broadcast transmitters they are put in a special class by the Federal licensing authority and as such operate under special conditions as to power rating and plate efficiency.

In the transmitter under consideration (Fig. 246) low-level modulation is employed in the stage succeeding the second amplifier. This may be either grid or plate modulation, as desired.

115. The Modulating Amplifier—Push-pull System. *Grid Modulation.*—A schematic diagram for a grid-bias modulating system is given in Fig. 123, which is a balanced push-pull amplifier. The radio-fre-

quency input to terminals 5, 6, and 7 is obtained from the preceding stage at corresponding terminals. This voltage is impressed upon the grids of the vacuum tubes out of phase; that is, when one grid is positive the other is negative and the amplified radio-frequency voltage appears across the plates similarly but in reverse order. The output of the tubes is delivered to the coupling transformer T_2 which must be tuned to the carrier frequency by the mid-tapped condensers C_5 and C_6 . The secondary of the transformer is tuned by condenser C_7 and delivers the radio-frequency power to the load resistance R_1 and to the input of the succeeding stage, the first power amplifier. Thus far, the circuit is simply a push-pull radio-frequency amplifier, with the grid bias adjusted to the I_p cut-off point or beyond. The circuit must be neutralized or

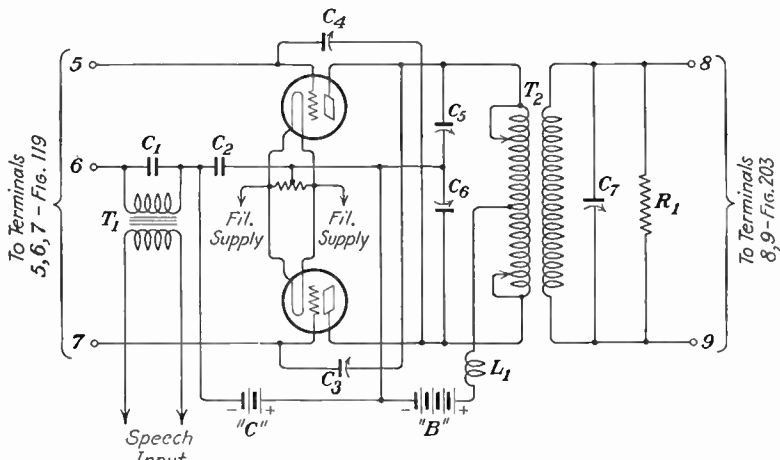


Fig. 123.—Circuit of push-pull type of modulated amplifier (grid modulation).

balanced to prevent self-oscillations, and this is done by means of the cross-connected balancing condensers C_3 and C_4 which, when properly adjusted, have approximately the same capacity each as the grid-plate capacity of each vacuum tube. These condensers feed back to the grids a voltage equal to, but opposite in phase to, the voltage fed back from the plate circuit by the grid-plate capacity, thus preventing self-oscillations.

When modulation occurs in this stage; that is, when audio-frequency voltages are induced in the secondary of the audio transformer from the speech amplifier, they are impressed upon the grids in phase, both grids being either positive or negative at the same time. But this in-phase condition is the same as that previously existing as far as the grid bias is concerned, hence the audio voltages act with the grid bias, thereby making it larger or smaller depending upon which half of the cycle is effective at the particular moment. Thus the total grid bias is made to vary in accordance with the audio input, and, following the normal vacuum-tube action of grid voltage upon plate current, the plate current is made to increase when the total negative bias is low and to decrease when the bias is high. These changes of plate current produce corre-

spending changes in the radio-frequency output, which is the effect explained under amplitude modulation. Therefore, the audio frequency causes a change of the bias, a change of the plate current, and resulting variations of radio-frequency output corresponding to the variations in the audio-frequency input voltages—hence the term *grid-bias modulation*. One of the advantages of this system is that, although the modulating stage may be handling considerable power, very little audio-frequency power is required to modulate it.

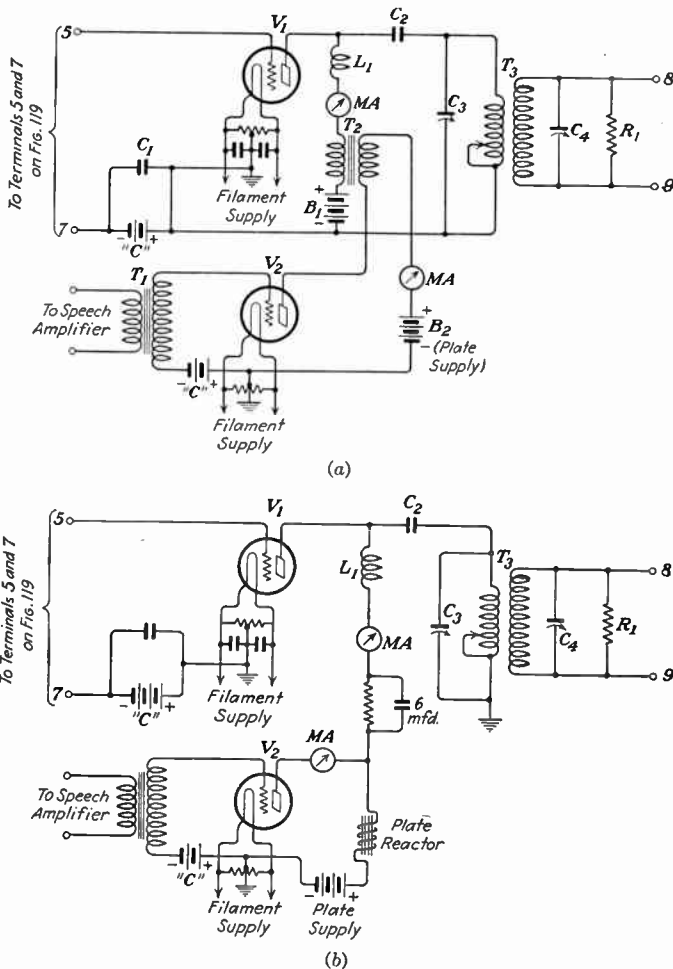


FIG. 124.—Heising systems of modulation: (a) modified system utilizing transformer coupling; (b) basic system utilizing reactor coupling.

116. The Heising Modulating Systems.—Figure 124(a) and (b) show circuits for the Heising systems of plate modulation. In these circuits, the vacuum tube V_1 amplifies the radio frequency applied to its grid and filament from terminals 5 and 7 which are connected to correspondingly numbered terminals (the output) of the second amplifier.

Condenser C_1 is a by-pass for the high frequency around the biasing potential, and C_2 is a blocking condenser to prevent short-circuiting the plate-supply battery B_1 . Condensers C_3 and C_4 tune the output transformer T_3 , the secondary of which delivers the radio-frequency power from the tube to the load resistance R_1 and the first power amplifier which is coupled to terminals 8 and 9. The tube V_1 is the modulating amplifier in the plate circuit of which the modulation occurs.

The vacuum tube V_2 is the modulator which is an audio-frequency amplifier. Speech or music from the preceding speech-amplifier system is impressed upon the input circuit of this tube by means of transformer T_1 . The amplified output is delivered to transformer T_2 the secondary of which is connected in series with the plate supply to V_1 . Inductance L_1 is a radio-frequency choke for isolating the plate-supply circuit as previously explained in the series- and parallel-feed systems. The modulator is designed to produce across the secondary of T_2 a voltage just equal to that of the plate voltage on V_1 supplied by battery B_1 , when the audio-input voltage to V_2 just reaches its maximum.

Suppose that the plate voltage on the tube V_1 [Fig. 124 (a)] is 1,000 volts and that without modulation the peak radio-frequency current in the load resistance R_1 is 1 amp. Then, let an audio voltage be applied to the input of V_2 , of such a wave form as shown in the lower wave of Fig. 121. Just when the audio wave starts at point B on this curve, there is no change of plate current in V_2 , hence no voltage is built up across the secondary of T_2 . But as the audio voltage increases to the point C on the curve, the secondary voltage of T_2 increases, finally reaching a maximum of 1,000 volts. Assume that for this half of the audio wave the secondary voltage is in series, adding to the battery potential of B_1 , that is, the upper side of the secondary is positive, thus adding to the voltage of B_1 . The total voltage on the plate of V_1 is now 2,000 volts. With double the plate voltage, the plate current doubles and the peak radio-frequency current in R_1 is doubled to 2 amp. Conversely, on the other half of the audio wave, from D to F on the curve of Fig. 121, a secondary voltage will be built up in T_2 and be equal to 1,000 volts at the point E , but this voltage will be negative and will oppose that of B_1 . The plate voltage on V_1 now totals zero, hence the plate-current and radio-frequency output-current peaks will be zero. At the point F on the audio wave the voltage to V_2 is zero, there is no voltage induced in T_2 , and the tube V_1 operates normally with 1,000 volts on the plate. Thus, in passing through the entire audio cycle, the plate voltage on V_1 effectively varies from normal, up to twice normal, back to normal, down to zero, and finally back to normal, completing the cycle. Accordingly the carrier-frequency peak currents are caused to vary correspondingly, that is, the amplitude of the carrier wave is made to vary in accordance with the audio wave. This is called *amplitude modulation*.

The circuit illustrated in Fig. 124 (b) is the basic Heising plate-modulation or constant-current system of modulation.

In this arrangement both the modulator and the modulating radio-frequency amplifier tubes, V_2 and V_1 , respectively, receive their plate supply from a common source through a large choke coil or plate reactor. The successful operation of this system depends almost entirely upon the design of the reactor choke coil as may be seen from the following explanation.

Let it be assumed that a steady plate current is flowing in the respective plate circuits of V_1 and V_2 and that a radio-frequency voltage is being developed across the

Therefore,

$$Z = \frac{X_L^2}{R} = \frac{1,000^2}{50} = 20,000 \text{ ohms}$$

If it is desired to make this impedance match the plate resistance of the tube, either the number of turns on the coil are changed, thus changing the upper term of the impedance equation, or the resistance R_1 is adjusted to vary the lower term of the equation. If the coil turns are increased the impedance increases, and if the resistance is increased the impedance decreases.

The output circuits employed in the various amplifiers of the radio transmitter of Fig. 104(c) are not simple single coils but coupling transformers. Figure 126 illustrates such a circuit. The transformer consists of the primary L_1 and the secondary L_2 , with the mutual inductance between them represented by the symbol M . The primary is tuned by condenser C_1 , while the secondary is tuned to series resonance by condenser C_2 . R_2 is the load resistance in which the radio-frequency power is dissipated.

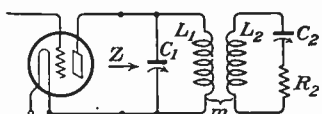


FIG. 126.—Amplifier coupled to its load through a coupling transformer.

With the circuits tuned as indicated above, the entire secondary circuit may be reduced to the equivalent shown in Fig. 125. It is possible to theoretically replace the secondary circuit with its load resistance R_2 by the primary inductance L_1 and an equivalent resistance R_1 in series with it. This method of dealing with the complex transformer is used merely to simplify the calculations of impedance and other functions, and the process of theoretically substituting the equivalent resistance R_1 for the true load resistance R_2 is called *reflecting the resistance* into the primary circuit. What actually happens is that, owing to the action of the transformer, the current flowing through R_2 is of such a value as to cause a voltage to be induced back into the primary circuit, of such value and phase angle that this reinduced voltage is equivalent to the voltage drop which would occur in a resistance of the proper value if it really were there. The numerical value of the reflected resistance is equal to

$$R_1 = \frac{X_M^2}{R_2}$$

where X_M is the mutual reactance of the primary and secondary of the transformer, and R_2 is the load resistance (the resistance of L_2 is assumed to be very small and is neglected here).

To calculate the impedance of the new equivalent circuit, the same procedure and formulas are used as for Fig. 125, except that the lower term in the impedance formula is now the reflected resistance, plus the coil resistance of L_1 .

In the radio transmitter of Fig. 104(c) the buffer amplifier employs the fixed tuned transformer so its impedance is not adjustable. The

second amplifier transformer is adjusted to obtain a relatively high impedance for voltage and power amplification by keeping the inductance large and the reflected resistance small. In the modulating stage the output impedance is adjusted to obtain linear amplification as explained in Sec. 84. This impedance may be found to be necessarily equal to the plate resistance of the tube, or some value greater than that, up to twice the plate resistance.

118. The first and second power amplifiers, as their names imply, are purely power amplifiers designed to bring the power level of the output of the modulating amplifier up to the value required for the maximum rated output of the radio transmitter. The circuits for these stages may be of the general form shown in Fig. 123, except that the speech-input transformer T_1 would be eliminated. Each of these stages handles the modulated carrier and, therefore, must be just as carefully tuned and adjusted for proper output impedance as the modulating stage.

119. Power-amplifier Grid Excitation.—The calculation of the proper amount of grid excitation necessary to drive a power-amplifier tube efficiently is a rather involved procedure since it must take into account the type of tube used and the load impedance into which the power-amplifier tube must function. Various values of grid excitation may be used for a given tube, depending upon the particular value of load impedance required, to develop a given amount of power across the load impedance. *A fairly accurate ratio of the power output necessary in the driver tube to excite properly a power-amplifier tube is that the excitation be approximately 10 per cent of the amplifier-power output required.* That is, if a power-amplifier tube is to deliver 500 watts into its load circuit, it must receive a grid-excitation power of approximately 10 per cent of this value, or 50 watts. Thus the driver tube must be capable of delivering to the grid of the power-amplifier tube at least 50 watts for proper grid excitation.

Although this ratio is fairly accurate in general practice, a more precise method would be to calculate the input grid voltages, plot a graph for different values of load impedances, and note the various output effects as to linearity and power until the desired combination is obtained.

In the power amplifier illustrated in Fig. 131 the proper value of grid excitation is regulated by the shunt resistances connected between grid and filament of the two power-amplifier tubes.

If the two power tubes are to deliver an output of 500 watts to the load circuit, the input power supplied to the grids is generally 50 watts. The resistance may then be regulated until the desired input drive is obtained, as will be explained later.

Another method of determining the grid-excitation voltage or power input necessary is to place two radio-frequency ammeters in the tank

circuit of the driver tube. If these two meters show identical values, the voltage

$$E_g = IX_L \tag{111}$$

This arrangement is illustrated in Fig. 127.

Another simple method of determining grid excitation is illustrated in Fig. 128, in which the grid-excitation voltage E_g is determined by the product of the current flow, as indicated by the radio-frequency ammeter M , and by the reactance of the condenser C_1 .

For example, assuming that the tank circuit L, C_1, C_2 , has been adjusted to resonance as indicated by the meter M at a frequency of

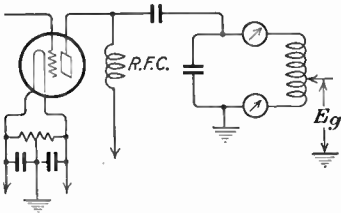


FIG. 127.—Radio-frequency ammeters in tank circuit to determine grid excitation necessary.

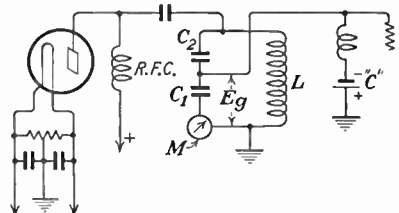


FIG. 128.—One radio-frequency ammeter M determines grid excitation in this circuit.

say, 1,000,000 cycles, and the condenser has a capacity of 0.01 mf. at this adjustment, the reactance

$$X_c = \frac{1,000,000}{2\pi fC} = 16 \text{ ohms (approx.)}$$

If the meter M reads 5 amp., then,

$$E_g = IX_c = 5 \times 16 = 80 \text{ volts}$$

Perhaps the simplest and most accurate method is that illustrated in Fig. 131 which is a modification of the arrangement used in the W. E. 12A transmitter. The power is developed in the load resistor R . Let us assume that this power must be sufficient to drive two 1-kw. tubes such as the W. E. 279A type, the unmodulated or carrier frequency of which delivers an output of 1 kw. into the load circuit L_3, C_3 . These particular tubes are rated at 1 kw. but are capable of a peak load as high as 2 kw. without overloading. That is to say, if the carrier frequency is adjusted to 1,000 watts of power, each tube dissipates approximately 500 watts of power in a balanced or push-pull arrangement as shown. Since the modulated peak power will be four times the carrier power at 100 per cent modulation, or 4,000 watts, these tubes must be and are capable of accommodating these swings without overloading.

However, in the adjustment of the transmitter the excitation power and the output power are based upon the carrier or unmodulated power to be used, which in this circuit is 1,000 watts.

Before calculating the necessary drive let us examine the characteristic curves of this particular tube (W. E. 279A) to ascertain the proper operating condition. Figure 129 illustrates the static I_p - E_c characteristic curve at the normal-operating plate potential of 3,000 volts and a bias potential of 275 volts. Figure 130 illustrates a dynamic characteristic curve which shows the amount of power that can be obtained in the load circuit with and without the flow of grid current.

It is important in any power amplifier to maintain the grid-current flow as close to zero as possible because, if the modulated radio-frequency peak voltages applied

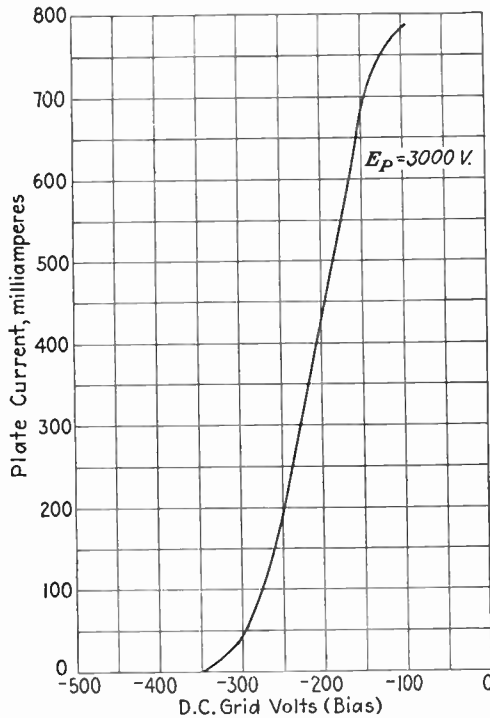


FIG. 129.—Static characteristic curve of W.E. 279A vacuum tube. (Courtesy of Western Electric Company.)

to the grid of the power tubes exceed the value of the grid bias, a grid current will flow during the applied positive peaks. This may result in serious audio-frequency distortion owing to the fact that the grid-current flow will be 180 degrees out of phase with the plate current. In view of the fact, however, that, at the 100 per cent modulation in the usual program transmission, only occasional peaks reach the high-percentage point, the positive grid current may go as high as 0.2 ma. without noticeable audio-frequency distortion. As a matter of fact, a very slight flow of grid current during the positive intervals of complete modulation will tend to increase the plate efficiency of the tube and consequently enable it to handle more power.

It is a very simple matter to determine if the grid current is excessive by inserting a direct-current ammeter or milliammeter in the tube plate circuit. If, during modu-

lation, the meter fluctuates excessively, it is probable that the amplifier is distorting. In other words, an excessive grid current acts as a load upon the plate current and owing to its 180-degree phase difference, alters the true shape of the plate-circuit component. This condition may be due to excessive grid excitation, improper plate and grid-bias voltages or improper load-impedance relations.

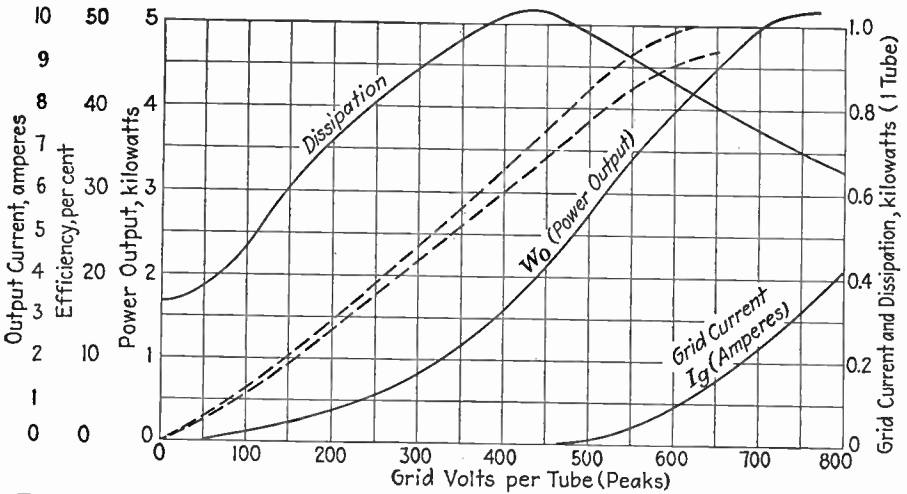


FIG. 130.—Dynamic characteristic curves of two W.E. 279A vacuum tubes in push-pull operation. (Courtesy of Western Electric Company.)

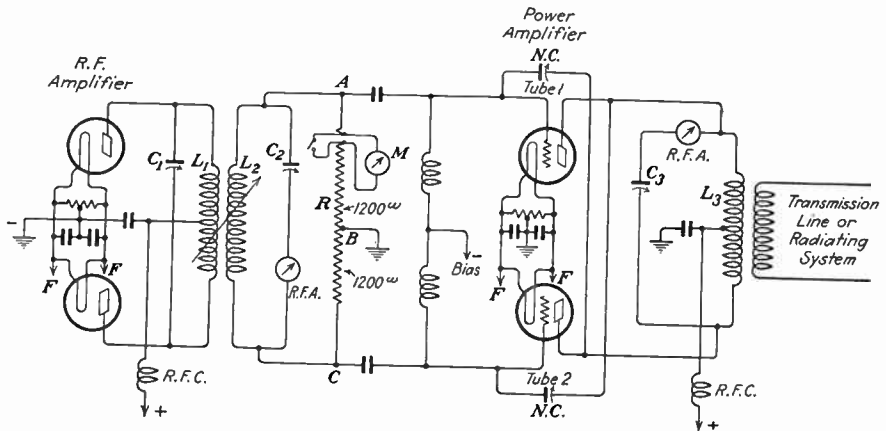


FIG. 131.—Balanced or push-pull linear amplifier.

Now let us examine the 279A tube *dynamic* characteristic for a specified frequency and load impedance, and ascertain what value of grid-excitation voltage will be necessary to excite it properly with a minimum of grid-current flow and a maximum power output. Figure 130 illustrates the approximate characteristics for this tube at the rated voltages when functioning into a load impedance of about 3,500 ohms.

In a balanced or push-pull linear amplifier the tuned-load impedance due to L_3C_3 (Fig. 131) is adjusted to four times the value of the plate resistance of a single tube in order that each tube will function into an impedance at least equal to its own value so that the proper power dissipation into the load will be obtained (see formula 110).

The top dotted curve in Fig. 130 illustrates the approximate efficiency of the tubes with relation to the excitation voltage and the lower dotted curve shows the value of the tank power in the load circuit L_3C_3 . It will be noted that the two curves are practically linear over a grid-excitation voltage ranging from 0 to approximately 650 volts. These curves assume a frequency of 1,000 kc./s. and a load impedance of 3,400 ohms.

Now, if the tubes are biased almost to the point of plate-current cut-off (class B), which in this circuit is produced by a negative bias of approximately 275 volts on the grid at a plate potential of slightly over 3,000 volts, there will be produced in the plate load circuit a power output representing the square of the input-grid voltage. The peak output will be four times greater than the unmodulated carrier power when 100 per cent modulation occurs. The proper value of grid-excitation voltage must, therefore, be determined in order that the unmodulated carrier-frequency power in the plate load circuit will be 1,000 watts. Consequently each tube must deliver 500 watts into the load circuit.

To accomplish this, grid excitation of about 265 volts must be applied to each tube, or, in other words, a potential of 530 volts across the two grids at the points *A* and *C* of the excitation resistance *R* in Fig. 131.

To develop an e.m.f. of approximately 530 volts across *R*, the entire resistance must dissipate 100 watts. To adjust for this, a radio-frequency ammeter *M* is connected in series with the resistance *R* and the radio-frequency drive varied until the meter reads a value of current which, when squared and multiplied by the resistance, will equal about 100 watts. An indication of about 0.22 amp. will develop 100 watts in the 2,400 ohms resistance. Thus

$$\begin{aligned}W &= I^2R = 116 \text{ watts} \\E &= IR = 528 \text{ volts}\end{aligned}$$

which is close enough to the required value.

All of these calculations are based upon the unmodulated carrier frequency and a load impedance of approximately 14,000 ohms looking into L_3C_3 at the assigned operating frequency. At 100 per cent modulation the peaks developed in the load circuit will be 4 kw. of modulated power or 2 kw. per tube. This power dissipation is readily accommodated by these tubes since it occurs only during occasional peaks. It is extremely important that this load impedance be correct for the type of tube used, not only for the proper development of power, but also for the ability of the tube to deliver a linear characteristic across the load. It will be remembered that if the load impedance is too high, the insufficient development of power across it will cause the modulation percentage to

be too low. This would result in a downward deflection of the output radio-frequency ammeter during modulation. If, on the other hand, the load impedance is too low, the excessive power developed will be dissipated in the tube and not the load, thereby causing an overheating and overloading of the tube itself. In addition, the modulation percentage due to the low-load impedance may be too high and consequently may produce serious audio-frequency distortion.

120. Calculation of Load Impedance in Power-amplifier Systems.—It has previously been shown that the maximum power output in a power-amplifying system is obtained when the load impedance is matched to the tube impedance, R_p , and the maximum undistorted power output, when the load impedance is twice that of the tube impedance. Although this is a true general statement the only definite assurance would be to plot a number of curves I_p - E_p against different values of grid bias E_c until an optimum value of load impedance is obtained. With this method it will not only be possible to obtain the desired load impedance for maximum power output but also the correct load impedance for minimum distortion (harmonic content).

Let us take, for example, the 10-type tube as a power amplifier. We see from the design characteristic that this tube has a plate resistance, R_p , of approximately 5,000 ohms at a plate potential of 400 volts. Under these conditions the plate current will be about 18 ma. at a grid bias of -30 volts. In accordance with general considerations of maximum undistorted power output, the load impedance should be double this value, or approximately 10,000 ohms. Now, since it must be assumed that we have sufficient input peak grid voltage or grid swing to drive the grids from the operating bias on the positive half of the input cycle to the point of zero bias and to twice the bias value on the negative half of the cycle, we may readily calculate the power output by taking the maximum and minimum values of plate voltage and plate currents and applying the ratio

$$P_o = \frac{(E_{max} - E_{min}) \times (I_{max} - I_{min})}{8} \quad \text{watts} \quad (112)$$

Now, since the excitation-grid voltage must not exceed the biasing voltage in order to keep the grid-current flow at a minimum, it is customary to calculate the excitation-grid voltage swing on the basis of an alternating-current-operated filament. That is to say, since the filament potential is of an alternating character, the bias must be so adjusted that the grid will at no instant in the cycle be positive with respect to the filament. This condition is assured by adjusting the grid bias so that the maximum grid swing will at no time exceed the biasing voltage. In an alternating-current-operated filament, therefore, the possibility of the grid bias being supplemented by the filament voltage is probable at some

instant of the input excitation cycle. To overcome this the excitation voltage is maintained below the biasing voltage by at least one half of the filament voltage. Thus, for the -10 tube with a filament voltage of 7.5, the excitation voltage must not exceed $30 - 3.75$, or 26.25 volts. In other words, the grid swing must be within the limits of -3.75 and -56.25 volts to prevent grid current from flowing.

For simplicity in calculation, however, we shall assume a steady direct-current potential applied to the filament so that we may also assume no change in grid bias will take place due to the filament supply. With this in mind, we shall proceed to calculate the proper load impedance by referring to the characteristic curve of a -10 tube in Fig. 132.

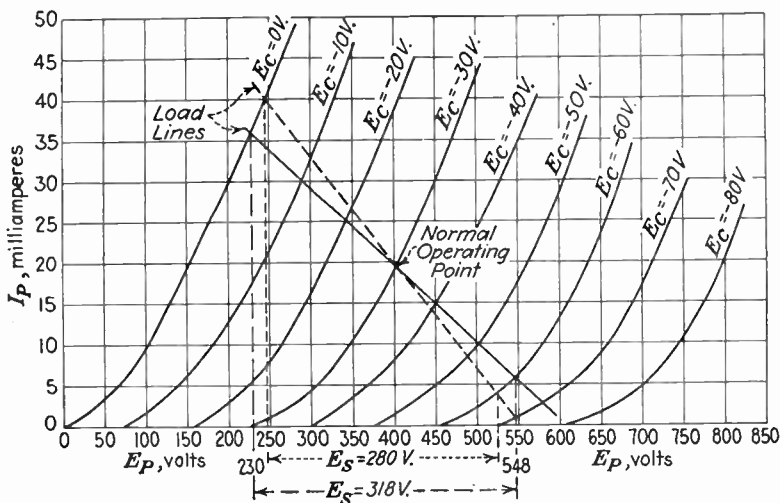


FIG. 132.—Load-line impedance determination, -10 tube.

Now, since the objective is to apply the proper grid swing (± 30 volts in this case) to the grid to vary it above and below the normal operating point (marked on the curve), so that when the grid swing is positive the plate current will rise to a value equivalent to the tube operating with zero bias, due to the neutralizing effect of the positive half-cycle against the fixed bias of -30 volts. In other words, this is the same as saying that an input-grid swing will vary the grid during its complete cycle from zero to twice the normal biasing value, or 60 volts.

If we draw a line from the normal operating point to intersect with the zero-bias curve at some arbitrary point, and also extend this line to the curve $E_c = -60$, we may readily determine the extent of the plate voltage swing E_s by ruling two lines perpendicularly to the plate voltage which corresponds to the intersection. Since the curve intersects at 230 and 548, the resultant plate swing at maximum grid excitation will be 318 volts. Now, since the normal value of plate-current flow is varying

during this swing from 36 to 6 ma., the load impedance may readily be calculated from the ratio

$$Z_o = \frac{E_s}{I_{p_{max}} - I_{p_{min}}} = \frac{318}{0.03} = 10,600^{\omega} \quad (113)$$

where E_s = the total plate swing between E_{max} and E_{min} .

Since this load impedance represents a value which is twice that of the tube impedance and, therefore, presents an ideal operating condition, we may proceed to calculate the power output and the harmonic distortion under this condition:

$$P_o = \frac{(548 - 230) \times (0.036 - 0.006)}{8} = 1,192 \text{ milliwatts (approx.)}$$

$$\text{Distortion } (H_2) = \frac{\frac{1}{2}(I_{max} + I_{min}) - I_{normal}}{I_{max} - I_{min}} \times 100 = 1 \text{ per cent} \quad (114)$$

Now, if on the other hand the dotted load line was chosen as our operating condition, we shall see that the output impedance output, power, and distortion would be as follows:

$$Z_o = \frac{280}{0.037} = 7,600^{\omega} \text{ (approx.)}$$

$$P_o = \frac{(525 - 245) \times (0.041 - 0.004)}{8} = 1,295 \text{ milliwatts (approx.)}$$

$$H_2 = \frac{\frac{1}{2}(0.041 + 0.004) - 0.018}{(0.041 - 0.004)} \times 100 = 12 \text{ per cent}$$

By comparing the two load lines it is seen that a greater power will be developed with the second line owing to the fact that its impedance value more closely approaches the plate resistance of the tube. However, we also see that the percentage of second-harmonic distortion is considerably higher, and, if we compute the relative power and harmonic increase of both load impedances, we shall see that by using the lower value of load impedance we obtain a power increase of approximately 7.8 per cent and a harmonic or distortion increase of approximately 91 per cent. It is therefore apparent that the slight increase in power obtained with a lower value of load impedance does not warrant its use in view of the large increase in the distortion content.

It may readily be seen that, if a number of load lines are plotted for a given tube, a desirable output impedance may be obtained in which the harmonic or distortion content may be reduced to a much greater degree than the power itself.

This method of calculating for the proper load impedance is applicable to all audio- or radio-frequency power-amplifier systems. When calculating the load impedance of a modulator tube, however, it is important that the plate swing E_s be double that of the radio-frequency

amplifier tube that is to be modulated (modulating radio-frequency amplifier). That is to say, if a given modulated amplifier tube is operated as class C at a plate potential of 2,000 volts, and, using the plate system of modulation Fig. 124(a) and (b), it will be necessary for the modulator tube to drive this voltage from zero to twice its normal value, or 4,000 volts, for 100 per cent modulation of the carrier frequency.

121. The Audio Monitor.—The purpose of the audio monitor connected to the output circuit of the second power amplifier in a broadcast transmitter is to demodulate a small portion of the radio-frequency power in the last stage of the transmitter. It then delivers the resultant speech or music to a loud-speaker, the input of which may easily be switched from the output of the monitor to the input of the speech-amplifier system, thus permitting a comparison of the quality of the program going into the radio transmitter with the program after modulation. Thus a constant check is available on the overall operation of the transmitter. Therefore, if the input quality of the program is satisfactory, and the output of the monitor sounds poorly, it is an indication that some part of the transmitter is out of order and a thorough check should be started at once to locate the difficulty.

122. The Frequency Monitor.—The purpose of the frequency monitor is to give the transmitter operator a constant check on the frequency of the emitted wave. This unit is connected to a very low voltage source in the output circuit and requires practically no carrier power from the transmitter. It is usually in continuous operation and requires very little attention other than to check the carrier frequency of the transmitter at regular intervals. Should the carrier show an appreciable deviation, such, for example, as 40 cycles, and a careful check indicates that the deviation is not in the frequency monitor itself, the carrier frequency may be corrected by adjusting the small trimmer condenser previously referred to as being across the quartz plate.

The operation of the frequency monitor is described in detail in Chap. IX on Broadcast Transmitters.

123. Antenna Coupling Circuits.—The method of coupling an intermediate-wave broadcast transmitter to the antenna is generally somewhat different from the methods used in short-wave transmitters, which will be described in the chapter on Antennas. It is more common to use direct coupling to the antenna in small intermediate-wave installations, and the losses due to the transmitter building being in the field of the antenna are not so serious as they would be on a short-wave transmitter. The circuit shown in Fig. 133 is a very common type of inductive coupling for broadcast frequencies. The capacity-coupled circuit shown in Fig. 134 has the advantage of greatly reducing all harmonics in the output owing to the low reactance of the coupling condenser to higher harmonic frequencies. In this circuit, C_1 is the tank condenser and C_2

is generally a mica condenser of large capacity compared to C_1 but is generally of a much lower voltage rating. This circuit may also be used with the point G_2 grounded and the antenna lead changed to position G_1 .

The transmission line connecting the transmitter with a remote antenna is used in high-power installation where it is desired to place the transmitter building out of the antenna field. The transmission line and the radiating system will be explained in Chap. VI on Antennas.

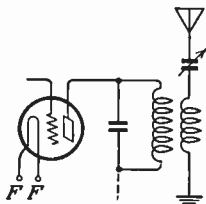


FIG. 133.—Inductive coupling for broadcast frequencies.

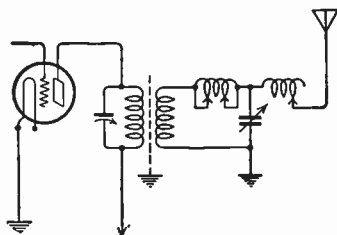


FIG. 134.—Capacitive coupling for broadcast frequencies.

124. Meters.—In all diagrams and discussions given, there has been no mention made of the location, number, or types of meters to be employed in the various circuits. In some circuits meters are always included, while in others only those absolutely essential are provided. It is of course desirable to have instruments in each circuit to assist in tuning the transmitter, to check, by means of current and voltage readings, all circuits during operation, and finally as a rapid method of locating trouble, should any occur. In general, the vacuum-tube oscillator requires a plate-current and grid-current meter. The buffer and second amplifier require only a plate-current meter, except in critical cases where excessive grid current may injure a tube. The modulator requires both grid- and plate-current instruments, to serve as a rough check on operation and distortion of the audio frequencies. All the remaining radio stages require grid- and plate-current meters and also thermocouple ammeters in the output circuits.

The last radio stage must have plate-current and plate-voltage meters and an antenna ammeter of an accuracy accepted by the Federal licensing authority. These instruments must be provided to permit checking the operating efficiency and output power of the radio transmitter. When a transmission line is used, a thermocouple ammeter is connected in the antenna lead and another instrument consisting of an external thermocouple and ammeter is connected in the ground lead. Two wires are connected to the external thermocouple and brought through a conduit to the transmitter building where connection is made to the indicating meter. Thus the thermocouple is located in the coupling unit, and the meter itself is located in the transmitter building, probably

on one of the panels of the transmitter, thus providing a means of constantly checking the current in the antenna-ground system by remote means.

125. The Power Supply.—Radio transmitters are classed, in regard to operating power, as being either direct-current operated or all-alternating-current operated. Ultimately the power source is generally alternating current, except on shipboard, and the difference of operation merely lies in the type of power-converting equipment used.

A transmitter which employs direct current throughout for filament, bias and plate supplies may be said to be direct-current operated, while if the filament supply is alternating current, operating heater-type tubes, or mid-tapped filament transformers and resistors, and if plate- and grid-bias voltages are supplied by a rectifier, the transmitter is classed as being all-alternating-current operated. It is also possible to operate the filaments on alternating current and to use a direct-current power source for bias and plate power. When this is done, the transmitter is not considered as being *all-alternating-current* operated.

The Filament Supply.—On direct-current-operated transmitters the filament supply is obtained from a direct-current generator usually driven by an alternating-current motor. On transmitters using alternating-current tubes, the filament power is obtained direct from the power lines through a step-down transformer. This transformer may have a common primary and several secondaries with different voltages available or, the several secondaries may be for the purpose of isolating certain tubes. A common secondary for all tubes may be used instead of separate secondaries if desired. As transformers require no maintenance, this method of providing filament power has some advantages over direct-current generator systems.

The Plate Supply.—To keep down extraneous noises and hums in the output of tubes, it is necessary to have a practically pure supply of direct current. This may be supplied either by motor-generator systems or by some type of rectifier. Both of these types of power supply are explained in later chapters (see Chaps. XIII and XIV).

126. Harmonics and Their Causes.—Harmonics in tube plate circuits are generally due to distortion of the plate-current wave shapes in the power stage of a radio-frequency amplifier. This is primarily due to the fact that, when an amplifier receives large grid-excitation voltages, the plate current is driven into the upper curved portion of the characteristic curve. If, therefore, a radio-frequency power-amplifier stage is adjusted to the point at which the grid-excitation voltage is maintained so that the plate-current rise does not extend into the upper curved area, the harmonic generation may be practically eliminated. Unfortunately, however, if the grid-excitation voltage due to the carrier frequency is maintained at these limits, no assurance can be given that the same

condition will prevail when the carrier is modulated; not unless the modulation percentage is extremely small, and then the output-power efficiency is so low that operation would be impractical.

In order to increase efficiency, therefore, the grid bias (E_c) is greatly increased to, or nearly to, the point of plate-current cut-off when using class B amplification. When using class C amplification this bias is increased to from one and one-half to twice the cut-off value. The grid-excitation voltage (E_g) may then be greatly increased to swing far up on the straight portion of the curve. In doing this, the plate-current component resembles approximately a half-wave sine curve with its associated harmonic-current components.

Current and voltage relations in a tube circuit, using a high grid bias for transmitting, affect harmonic generation. When the grid-excitation voltage is low, the plate current practically follows the exciting-grid voltage. This is particularly true if the tube possesses a high amplification factor because then the plate reactionary effects are very small.

If, on the other hand, however, the grid excitation is increased, the plate current will dip due to the decrease in plate potential below the exciting-grid voltage. This may be clearly seen by a careful examination of Fig. 135 in which the grid-excitation voltages are of varied values such as during the process of modulation.

It will be noted that during the intervals of low grid-excitation voltages the harmonic content will be very small, and that during the period of high grid-excitation voltages the harmonic content will be greatly increased. Hence, since the grid voltage is a function of the degree of modulation, the resultant harmonic content will depend upon the average effect between the low- and high-voltage grid swings. In other words, although a transmitter is said to be modulated at 100 per cent, in reality these peaks are only of short duration and it is not to be understood that 100 per cent modulation means that all peaks are of that value. This means then that the actual percentage of modulation is really the *average* between the low and high peaks during the usual radio program. Hence, the average percentage of modulation during program transmission may be considered as being of a low value, although it is imperative that occasional peaks should reach the full 100 per cent point. It is the effect of these occasional peaks reaching 100 per cent which classifies a transmitter as being modulated at this percentage, and not the average degree of modulation which may be lower in percentage. Actually, if every peak during a program were to swing the grid to the point of 100 per cent modulation, serious distortion, due to harmonic generation, would result.

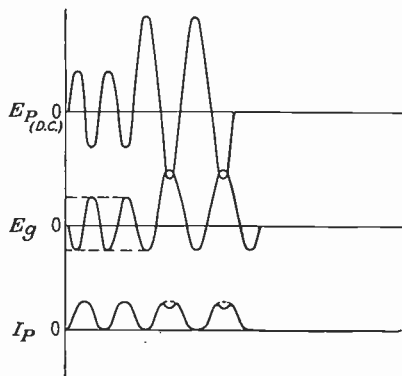


FIG. 135.—Current and voltage relations in a vacuum-tube circuit.

Thus it may readily be seen then that by the proper design of tubes and load impedances, together with the proper value of bias and plate voltages, the average harmonic content may be reduced sufficiently to

avoid distortion. Practically, this may be accomplished by reducing the plate-voltage component almost to the point in which the plate-current dip begins, as illustrated by the dotted lines on the I_p curves in the figure.

Thus it may be seen that the harmonics generated in tube circuits, due to the distortion of the wave shapes, may be greatly reduced by the proper use of the tubes, and the critical adjustments of the circuits into which they function. Unfortunately, however, the fulfilment of these conditions does not prevent the radiation of harmonics due to the antenna or improper coupling systems.

Since a transmission line or an antenna system has a distribution of inductance and capacity, some of the harmonic frequencies generated in the tube-output circuit may correspond with some of the harmonic frequencies of the transmission line or the radiating system and produce excessive harmonic radiation.

It is possible to determine the percentage of the harmonic-frequency content to that of the fundamental frequency by taking the ratio between the harmonic and fundamental fields created at a given distance from the antenna. That is, if the ratio of the two currents generated in the measuring equipment is taken and the radiation resistance of the fundamental and harmonic frequencies carefully measured, a definite relationship of the energy content of the harmonic frequency and the fundamental may be determined by the equation.¹

$$\frac{\epsilon(\text{harmonic})}{\epsilon(\text{fundamental})} = \frac{I_N(\text{harmonic current})}{I_A(\text{fundamental current})} \times \sqrt{\frac{R_H(\text{harmonic radiation resistance})}{R_a(\text{antenna radiation resistance})}} \quad (115)$$

If the harmonic contents are excessive, suitable means must be made to reduce them to negligible values.

127. The suppression of harmonics in transmitters is of the utmost importance to reduce the unnecessary interferences which result from this radiation in other frequency channels.

Harmonic radiation, particularly that due to the second harmonic of the fundamental frequency, is the most objectionable. Although this frequency cannot be entirely eliminated, several methods can be arranged for reducing its field intensity to a negligible degree. If the second-harmonic content at a distance of approximately 1 mile from the transmitter does not exceed 0.02 per cent of the fundamental frequency, the carrier frequency may be considered practically free of harmonic interference. In high-power transmitters the second-harmonic limit has been extended to 0.05 per cent of the fundamental.

¹ Labus, J. W. and Roder, Hans, The Suppression of Radio Frequency Harmonics in Transmitters, Proc. I.R.E., vol. 19, p. 949, June, 1931.

Several methods may be used to reduce the second-harmonic content of a wave by introducing low-impedance paths to ground at the harmonic frequency or by inserting a high-impedance series-resonant circuit in series with the radiating circuit to block out from the radiating system the undesirable harmonic frequencies.

Figure 136(a) illustrates the first method in which the two circuits are tuned to the second-harmonic frequency and thereby offer to the undesirable harmonics a low-reactance path to ground. The fundamental, or carrier, frequency, on the other hand, is unaffected, since the series-tuned circuits will offer a high reactance to it.

Figure 136(b) illustrates the second method in which the series-resonant circuit is tuned to the fundamental frequency. When the inductance L of the resonant circuit LC is adjusted so that the reactances X_L and X_C are of equal value, the total effective reactance will be zero. Thus, if this resonant system is placed between the power-amplifier output stage and the antenna radiating system, no reactance will be intro-

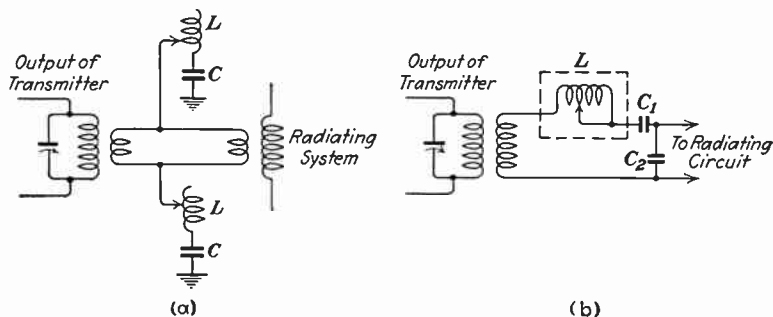


FIG. 136.—Harmonic reducing circuits.

duced to block the fundamental frequency, and the energy will be delivered practically unattenuated to the antenna system. The second-harmonic frequency, on the other hand, will not readily pass through the series-resonant circuit since the reactance of the circuit to this frequency is very high. The antenna coupling condenser C_2 is an important part of the harmonic suppression circuit since it acts as a reaction load across this circuit. The voltage developed across it, therefore, equals IX_c .

It has previously been shown that harmonics generated in the plate circuit of a class B or class C amplifier may be reduced by the proper adjustment of the tube voltages and load circuit. However, since the tank circuit may itself generate harmonics when set into oscillation, other preventive means must be used to overcome these harmonics. One of these, and perhaps one of the most extensively used, is the balanced, or push-pull, amplifier as illustrated in Fig. 203.

With this arrangement it is not only possible to obtain more than double the power output but also to eliminate all of the even harmonics to the transmission-line circuit through T_1 and across C_3 . This is due to the fact that the fundamental-frequency current flowing in each plate circuit is out of phase by 90 degrees. Hence, the even harmonic-frequency currents will flow through C_1 and C_2 in a direction so that the harmonics in the C_3 load circuit cancel.

This condition, however, is only true provided that the circuits are properly balanced; that is, the tubes should be well matched for plate-current flow for a given value of plate voltage and grid bias. The condensers C_1 and C_2 must be of the exact capacity to insure an equal value of reactance. All leads should be of the same length from the respective condenser and plate connections. If these conditions are not fulfilled, additional even harmonics may also be contributed to the transmission-line load circuit.

Accurate measurements of harmonic radiation can be made only with specially designed supersensitive equipment at some distance from the transmitter, but several simple methods may be used to enable the operator to determine approximately the ratio of the harmonic content to the fundamental frequency radiated.

One of these simple methods uses a precision-type wavemeter which may be placed near the output of the tube circuit to determine the presence of the second harmonic by tuning the wavemeter very accurately to the calibrated setting of the second-harmonic frequency of the carrier wave. If a radio-frequency ammeter is used to indicate this quantity, and a comparison made with the indication when the wavemeter is tuned to the fundamental frequency, an approximate percentage of the harmonic content may be obtained. The coupling between the wavemeter coil and the tank circuit, however, must not be altered, since this will affect the accuracy of the measurements. The current-indicating device must, therefore, have a sufficient range to cover the current readings for both the harmonic and the fundamental values to obtain a direct ratio. Circuit losses in the wavemeter at the higher frequency measurements will affect the accuracy to a slight degree but not sufficiently to prevent a fair comparison.

This same method may be more effectively used by placing the wavemeter at a fair distance from the antenna radiating system, as far removed from metallic structures as possible, and in this way determining if the harmonic content of the actual radiated wave is within the required percentage of the fundamental

128. Parasitic Oscillations.—The oscillation of radio-frequency circuits at frequencies other than that to which they are normally adjusted is generally referred to as *spurious* or *parasitic* oscillation. These oscillations may take place in neutralized radio-frequency amplifiers at extremely high frequencies owing to stray inductance and capacities in the connecting leads to the plate and grid circuits. This is particularly true in circuits where tubes are connected in parallel because of the paralleling grid and plate leads introducing small L and C oscillatory circuits.

The presence of parasitic oscillations may be readily detected by a large increase in the plate-current readings and an excessive overloading of the tube plates. If the plate circuits are not suitably protected by

overload circuit-breakers or fuses, the plates may become white-hot and ruin the tube. Choke coils and condensers may also be subjected to severe strain and damage, due to the high oscillating voltages built up across them.

Parasitic oscillations may be prevented by inserting very small radio-frequency choke coils or resistances in series at each one of the grid and plate connections. The chokes must have a very small inductance and should be capable of carrying the normal tube plate-current flow. A non-inductive resistance of approximately 50 ohms may be connected in each grid lead to suppress the radio-frequency grid-current flow generated by the parasitics. A non-inductive resistance may also be connected in each plate lead, but, since this circuit will carry the normal plate-current flow, the resistance would have to be of the heavy-duty variety to withstand the load. Choke coils are preferable, but they must be efficiently designed on Isolantite forms and possess negligible distributed capacity.

II. ULTRA-HIGH-FREQUENCY CIRCUITS

129. Ultra-high-frequency Systems. General Considerations.—Wave lengths below 10 meters (over 30,000 kilocycles) are generally termed *ultra-short-wave, ultra-high, or very high frequency emissions*. It is convenient when dealing with these frequencies to specify them in megacycles (mc. or mc./s.), the prefix *mega* meaning one million. As electric waves have been generated as short as 82 millionths of a meter and the gap between the radio and infra-red spectrums bridged, it is possible that practical apparatus utilizing these extremely high oscillations may ultimately come into use. Table XIV, entitled Wave Length—Frequency Conversion Table, which appears in the Appendix, will aid the student of ultra-high-frequency literature in understanding the relationship between frequencies expressed in megacycles (or even centimeters) and wave length in meters. Most practical apparatus for communication purposes operates in the spectrum between 5 and 10 meters, although on occasion shorter wave lengths have been used.

There are certain fundamental features of ultra-high-frequency circuits which make them different from circuits operating on the conventional lower frequencies. Distributed capacity and inductance take on an importance not so noticeable with lower frequencies. Component parts of the circuit must be built with low-loss parts, and mechanical rigidity of the circuit becomes of great importance as any change in relative physical position of circuit parts causes a frequency change with consequent variation in output.

Owing to the very high frequencies involved, skin effect and other circuit losses require the use of conductors offering large surface paths for the currents. Careful consideration must, therefore, be given to

apparatus arrangement, and leads must be kept very short or eliminated wherever possible.

When the very high frequencies are to be used, the inter-element travel time of the electrons in the tube becomes important. It has been found that tubes especially designed for ultra-high-frequency generation must have electrodes of comparatively small area and a structure short in the axial direction.

Frequency control of ultra-high-frequency generators by means of quartz crystals becomes impractical because of the frequency multiplication necessary and the extra tubes and circuits necessary for doubling and triplication. Considerable success has been had in Germany with tourmaline crystals¹ which can be ground to a frequency of such a high value that frequency doubling or triplication is unnecessary. Until frequency stabilization on the ultra-high frequencies becomes necessary, crystal control will probably not be used to any great extent. The disadvantages of crystal-control design are the high cost of suitable crystals, the fact that a separate crystal is usually necessary for each working frequency (unless harmonic operation is possible), the selective receiver necessary if the benefit of crystal control of the transmitter is to be secured, and the considerable skill on the part of the operator which becomes necessary for good results.

130. Ultra-high-frequency Oscillators. *a. Regenerative Oscillators.*—These oscillators, as used for practical radio purposes, are divided into two groups: (a) regenerative oscillators, (b) electron oscillators. Among the first of the practical ultra-high-frequency vacuum-tube oscillators was the arrangement developed by Gutton and Touly about 1919. This oscillator employed a very simple circuit which consisted of a single turn of inductance between the plate and the grid of the tube, with a grid condenser and grid leak, arranged as shown in Fig. 137(a). This circuit tuned down to 2 meters and is the fundamental-circuit basis for most of the single-tube regenerative ultra-high-frequency oscillators used at the present time.

Another widely used fundamental circuit is that developed by Mesney in 1924 and which is shown in Fig. 137(b). This is a two-tube balanced circuit and gives stable oscillations down to 1.5 meters.

For wave lengths between 5 and 10 meters the conventional tuned-plate tuned-grid oscillator is commonly used. This oscillator when used with a radio-frequency power amplifier is usually capacity coupled to the amplifier.

The principal difficulties with ultra-high-frequency oscillators are caused by the inter-electrode capacity of the tube or tubes being too high

¹ See STRAUBEL, H., Fundamental Crystal Control for Ultra-high Frequencies, *QST*, April, 1932, p. 10.

and because it is impossible to get proper impedance values with the low L and C used in the tuned circuits. It must also be remembered that when a tube oscillates, electrons pass to-and-fro within the tube. As the frequencies become higher, the length of time it takes for the electrons to complete their inter-electrode journeys becomes a limiting factor on the oscillating frequency.

Lieutenant Wenstrom, U.S.A., reports¹ the following practical results with ultra-short-wave oscillators of the regenerative type:

1. Carefully used, the Lecher system is a precise method of wave-length measurement (± 0.01 per cent).
2. Relative power output and efficiency of ultra-short-wave oscillators under various electrical conditions can be determined with about ± 3 per cent precision by means of a closely coupled high-resistance loop in series with a thermoammeter.

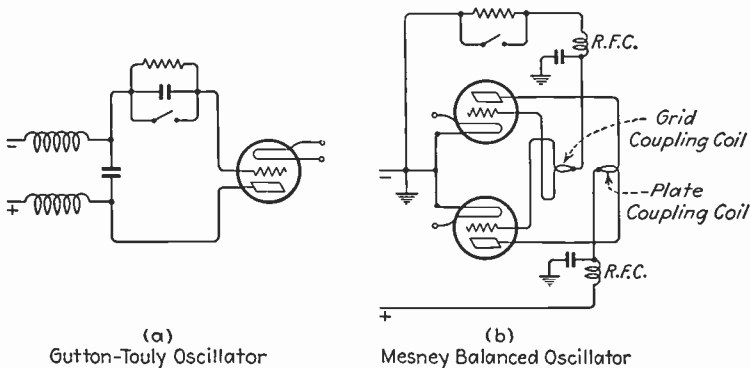


FIG. 137.—Ultra-high-frequency oscillators, regenerative types.

3. Absolute efficiency can be determined by the heat-radiation method with better than ± 2 per cent accuracy.
4. The absolute efficiency of ultra-short-wave oscillators operating under average optimum conditions is about from 30 to 40 per cent.
5. Operating with grid resistance, the single-tube oscillator is more efficient than the balanced oscillator at low or high plate voltage, while at moderate plate voltage the two circuits are about equal in efficiency.
6. Operating with grid battery, the balanced oscillator is more efficient than the single-tube oscillator at high negative grid bias, while with moderate grid bias the two circuits are about equal in efficiency.
7. In the optimum grid-resistance region (5,000 to 10,000 ohms per tube), a 2-to-1 variation of grid resistance has little effect on output or efficiency.
8. With zero grid resistance the efficiency of both types of oscillators is low.
9. Considerable changes in filament voltage (± 20 per cent) have little effect on output and efficiency of oscillators employing thoriated-filament tubes.
10. In properly designed circuits the filament-emission demands are not far from normal.

¹ This list of Conclusions is taken from the paper, "An Experimental Study of Regenerative Ultra-short-wave Oscillators," by William H. Wenstrom, *Proc. I.R.E.*, vol. 20, no. 1, January, 1932.

11. The variations of oscillating current, plate current, and grid current in ultra-short-wave oscillators are less than 5 per cent during an hour of continuous operation.

12. The wave length of an ultra-short-wave oscillator varies with time after plate voltage is applied due to the plate-grid-capacity change caused by heating.

13. The wave length of a balanced oscillator in general increases with time at a decreasing rate. The change is about 0.01 per cent per minute after 8 min. operation.

14. The wave length of a single-tube oscillator at first decreases with time owing to the rapid capacity change in the external circuit condenser, then increases with time owing to the predominant plate-grid-capacity effect.

15. The wave length of either type of oscillator varies only slightly with changes in filament voltage. It varies markedly, however, in response to plate- or grid-voltage changes.

An entirely different method of producing ultra-high-frequency oscillations was discovered by Barkhausen and Kurz and reported by them in 1920. They called this circuit an electron oscillator and its

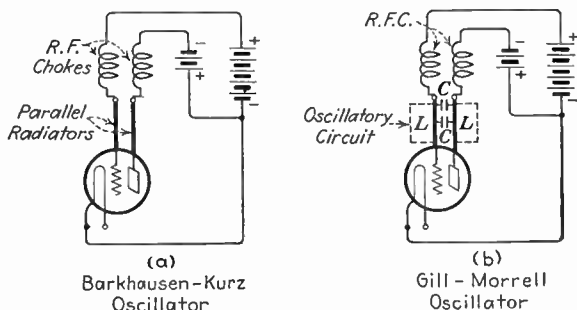


FIG. 138.—Ultra-high-frequency electron oscillators.

principal difference, in so far as circuit is concerned, is that the grid is maintained at a high positive potential while the plate is at a zero or slightly negative potential. The theoretical operation is as follows.

b. Electron Oscillators.—In the Barkhausen-Kurz oscillator system, shown in Fig. 138, the oscillations are due entirely to the to-and-fro movement of electrons between the grid and filament of the vacuum tube. This is to say, the oscillating condition of the electrons takes place solely within the tube and is not determined by the circuit constants as in the conventional regenerative oscillator.

The oscillating electrons are a result of a high positive potential being applied to the grid and a small negative potential to the plate. As the electrons are accelerated toward the grid at high speeds, they pass beyond the grid lateral structure toward the plate but are immediately pushed and pulled back to the grid. This movement of electrons back and forth through the grid structure apparently causes a displacement of the free electrons in the grid wires, first in one direction and then in the other, resulting in an oscillating potential change between the grid and the plate of the tube. Hence, if these oscillations are of an amplitude great enough to develop sufficient power to overcome the circuit losses, the oscillations will be sustained.

The oscillations produced by the oscillations of electrons within the tube are called *Barkhausen-Kurz* (or B-K for short) *oscillations* and the inventors of the circuit claimed that the frequency of oscillation within

the tube was not affected by adjustment of any external circuit connected to the tube. Later discoveries by Gill and Morrell showed that, if an external circuit was connected to a B-K oscillator, it would under certain conditions oscillate too, and that the frequency of these oscillations which then took control of the circuit was the frequency of the tuning system and they were effected by the adjustment of the external connections. These new oscillations are known as *Gill-Morrell* (G-M) *oscillations* and it has been shown that whether the B-K circuit produces B-K or G-M oscillations depends upon the circuit adjustments. The Gill-Morrell oscillator is in fact a Barkhausen-Kurz circuit in which the external circuit connected to the grid and plate is a certain critical value which produces G-M oscillations. A slight readjustment of this circuit may throw the oscillator back to B-K characteristics. In other words, whether an electron oscillator is B-K or G-M is only a matter of circuit adjustment which is very critical in nature.

The formula¹ for wave length for the B-K oscillator as given by the inventors of the circuit is

$$\lambda = \frac{1,000(d_a E_g - d_g E_a)}{\sqrt{E_g(E_g - E_a)}} \quad (116)$$

where E_a and E_g are the respective anode and grid potentials, and d_a and d_g are their respective diameters.

Both of these oscillators are capable of producing extremely high frequencies, in the order of 300,000 kilocycles (1 meter) and up but are limited to very small powers.

Specially designed tubes are required for these oscillators to keep the efficiency extremely high during the period of oscillation, otherwise the oscillations are of slight duration.

131. Modulation of Ultra-high Frequencies.—For the regenerative type of oscillator the Heising constant-current system is very satisfactory and can be applied to high-level modulation on M.O.P.A. circuits.

A considerable study of modulation of B-K oscillators was made by Hollmann who reported, in 1928, several schemes for accomplishing such modulation as follows:¹

a. In one, the secondary of the microphone transformer was connected in the anode (negatively biased) lead. In another, the anode of the modulator tube was connected to the grid (high +) of the oscillator.

b. With a receiver similar to the oscillator but having headphones connected in series with the anode (negative bias), circuit telephone transmission was accomplished up to a distance of 500 wave lengths.

¹ WENSTROM, WILLIAM H., Historical review of Ultra-short-wave Progress, *Proc. I.R.E.*, vol. 20, no. 1, January, 1932.

Owing to the relatively greater permissible width of the operating spectrum employed with ultra-high-frequency emissions high modulation frequencies are also permissible which is of advantage under certain conditions of service.

Note on Ultra-high-frequency Wave Propagation.—This subject is covered in Chap. VI on Antennas and Wave Propagation.

CHAPTER V

RECEIVING-CIRCUIT PRINCIPLES

The radio-frequency amplifier and the vacuum-tube oscillator described in the preceding chapter are used in both transmitting and receiving circuits. Regardless of the circuit in which they are used their theory of operation is the same. Therefore, these applications of the vacuum tube in receivers will not be further explained. The remaining features of the receiving circuit may now be logically studied.

132. General Receiving Considerations.—The radio receiver is a very important part of any radio system, and when it is designed it must be considered as part of a definite system.

All radio receivers have certain features in common, such as sensitivity and selectivity. On the other hand, certain classes of receivers have characteristics not required in other types. For example, a broadcast receiver must have fidelity (quality) in addition to sensitivity and selectivity. This factor is not required of receivers used for *CW* telegraphy and not so much so for speech-service receivers such as are commonly used in aeronautical and police work. Also it is undesirable that receivers feed energy into the antenna. These and other problems will be treated in this chapter.

I. THE DETECTOR AND AUDIO AMPLIFIER

133. The Screen-grid Tube.—Figure 139 shows the electrode arrangements (*a*) of screen-grid tubes and the inter-electrode capacity (*b*) between the various elements.

In the ordinary three-element vacuum tube there is existent a considerable amount of inter-electrode capacity between the grid and the plate which stores up a heavy electrostatic field between these elements.

The presence of this capacity between the grid and the plate does not only decrease the amplifying efficiency of the tube, due to the increase in space charge, but also introduces circuit reactions between the plate and grid circuits. This is particularly objectionable in radio-frequency amplifying circuits because of the possibility that feed back through the inter-electrode capacity may set the tube into regeneration or oscillation. Radio-frequency amplifiers must not go into self-oscillation if circuit stability and minimum distortion are to be maintained. Self-oscillation may be prevented by a neutralization system in which the internal

capacity effect of the tube is balanced out (see Balance Circuits, Sec. 86, and Neutralization of a Transmitting Circuit, Sec. 87).

In radio-frequency amplifiers where the screen-grid tube is used, this feed-back possibility is, however, greatly reduced because of the reduction in the inter-electrode grid-to-plate capacity caused by the interpolation of the screen grid between these two elements. This reduces the effective capacity between the plate and grid C_1 and C_2 in Fig. 139 by the same reasoning that connecting condensers in series reduce their effective capacity. The reduction of space charge may also be visualized as due to the neutralization of the negative field near the plate, because of the positive potential applied to the screen grid, thereby allowing a greater effect upon the plate circuit for smaller control-grid or signal voltages.

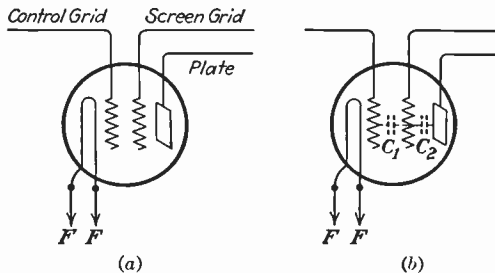


FIG. 139.—(a) Screen-grid tube (tetrode); (b) C_1 and C_2 represent interelectrode capacities.

Hence, when the effective capacity in any circuit is reduced, the capacitive reactance in ohms is increased. Thus, the capacitive coupling between the plate and control grid is greatly reduced when using the screen-grid tube, owing to its lower internal grid-plate capacitance. Also, in view of the greatly reduced electrostatic field within the tube, the amplification factor, or μ , of the tube will be greatly increased, thereby increasing the efficiency of the tube as a voltage amplifier in radio-frequency and audio-frequency circuits. Incidentally, the alternating-current plate resistance R_p of the screen-grid tube will also be greatly increased by the reduction in tube capacitance which will of course reduce the ability of the tube to be effective as a power amplifier. Screen-grid tubes, however, are usually used as voltage or stabilizing amplifiers and are not generally used in output or power circuits where any considerable amount of power dissipation is required.

It is of course very important, that the load impedance of a screen-grid tube must be at least equal to the tube impedance if a proper transfer of energy is to be expected. For example, if an ordinary triode, having a plate resistance R_p of, say, 10,000 ohms, functioning efficiently into its matched load of an approximately equivalent resistance, is replaced by a tube of the screen-grid variety, having a plate resistance of approximately 500,000 ohms, no appreciable increase in amplification can be expected unless the load impedance is increased to effect an approximately matched

condition. This condition may be easily arranged for in radio-frequency amplifiers by simply adding more turns to the load inductance to bring up the inductive reactance. In low-frequency or audio amplifiers, however, this would be considerably more difficult because the low frequencies between 30 and 200 cycles would not produce as much reactance across the load impedance as the higher audio frequencies, unless the units were designed to have a very large inductance so that when the lowest audio frequency is applied it will effect a reactance which will be an approximate match to the tube-plate resistance. This would require a specially designed transformer of large physical dimensions. In general practice where screen-grid tubes are used as detectors, they are usually coupled into a resistance-coupled audio-frequency amplifier where it is a relatively simple matter to match the load to the tube impedance by using a plate-coupling resistance of a value somewhere between 250,000 and 500,000 ohms (see Secs. 139*d*, and 218).

Variable- μ Tube.—This type of tube is of the typical screen-grid variety as illustrated in Fig. 139. The name *variable- μ* has been applied to this tube owing to its ability to control amplification by varying the voltage upon its control grid. That is to say, the structure has been so designed that the cut-off point of the I_p - E_c characteristic curve is such that a considerably higher grid-bias voltage is required to produce cut-off, relative to that in the ordinary screen-grid tube.

This may be more clearly seen if the relative I_p - E_c characteristic curves of both tubes are examined in Fig. 140. It will be noted that the I_p - E_c curvature of the ordinary screen-grid tube in (a) is more pronounced than that in (b) of the variable- μ tube. This indicates that, if an excessive grid swing is applied to the ordinary screen-grid tube, the asymmetry due to the lower curved area will produce radio-frequency distortion. In the variable- μ tube the approach to the cut-off point is more gradual, and consequently much more signal can be applied before distortion results.

Furthermore, the minimizing of the steepness at the lower portion of the characteristic curve decreases the possibility of distortion due to the intermodulation of the desired signal with that of an interfering signal. This is commonly known as *cross-talking* or *cross-modulation interference*.

134. The Vacuum-tube Detector.—The purpose of a vacuum tube as a detector or demodulator is to rectify the incoming signal of high frequency

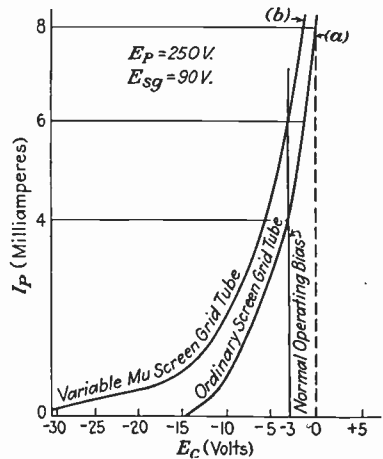


FIG. 140.— I_p - E_c characteristic of variable- μ (supercontrol) tube.

to one of low or audible frequency. This is accomplished by adjusting the vacuum-tube plate voltage and grid bias so that a symmetrical signal of high frequency may be rectified into asymmetrical waves or group pulses. This condition is necessary in order that the signals may pass through a high impedance such as a telephone or primary winding of an audio-frequency transformer. The two methods by which rectification may be obtained are known as *plate* detection and *grid* detection.

135. Plate Detection.—When an alternating e.m.f. is applied across the grid and filament as illustrated in Fig. 141 the normal or steady plate-current flow will be affected in accordance with the character of the potential change on the grid. That is to say, when the alternating potential is applied, a plate current *above* normal will flow when the grid is charged positive with respect to the filament, and a plate current *below* normal will flow when the grid is charged negative with respect to the

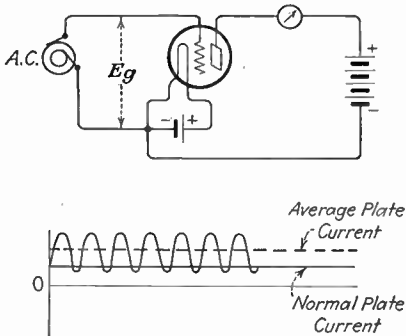


FIG. 141.—Triode plate-current variations.

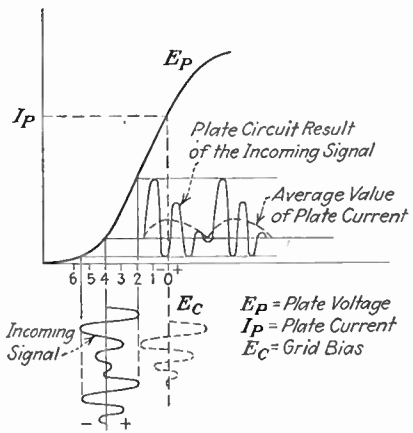


FIG. 142.—Graph illustrating rectification.

filament. The tube may be so operated that the increases in the plate current over the normal value may be much more than the decrease below the normal value. The average plate current, therefore, under these conditions is much more than the normal value. Consequently, if a signal variation of radio frequency is applied to the grid of the tube, the signal is not only amplified but also rectified. If, therefore, the signal is of a damped or modulated character, the plate-current variations will be asymmetrical in shape and thus make it possible for the audible groups or pulses to pass through the telephones or the primary winding of the audio-frequency transformer.

Figure 142 illustrates the condition in which a modulated signal wave is applied to the grid of a tube operating as a detector. The asymmetry, in this case, is fulfilled by the adjustment of the negative grid bias E_c , so that the signal is applied below the straight portion or knee of the characteristic curve. This produces a rectified component in the plate circuit which develops a voltage across the load impedance of

the telephone or primary winding. Since the voltage applied to the plate of the tube is maintained constant by the power supply, it can readily be seen that, when an incoming signal develops an IR or voltage drop across the load, the actual voltage on the plate of the tube will be varied accordingly. This variation may then be transferred to a succeeding or amplifying circuit through a coupling transformer or resistance.

In all radio-frequency detecting circuits, it is very important to build up a maximum audio-frequency voltage across the load impedance if the proper efficiency is to be obtained. In view of the fact that there are two components present in the plate circuit of a detector tube, namely, the carrier or radio frequency and the modulation or audible frequency, a small radio-frequency by-pass condenser must be shunted across the load impedance. This condenser must offer a high impedance to the modulated-frequency component in the plate circuit and a low impedance to the radio-frequency or carrier component. If this condenser offers too much reactance to the radio-frequency signal, some of the voltage will be spent in forcing current through the high-impedance primary load and consequently will result in a reduction of the signal voltage acting on the load.

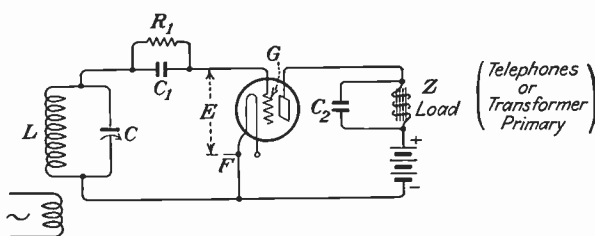


FIG. 143.—Grid-leak detector circuit.

Plate detection is generally used where signal voltages are of large magnitude such as is true when a number of stages of high-gain radio-frequency amplification precede the detector tube. If the signal voltage is small when operating a detector near the cut-off point of the I_p-E_c characteristic curve, the amplification will follow a square law and the result will not be linear. If, on the other hand, the signal voltages are of a larger magnitude, and plate detection is used, rectification will be fairly linear and better quality will result.

136. Grid Detection.—In the grid-detection method of rectifying or demodulating a radio signal the entire process of detecting takes place in the grid circuit, and the plate circuit merely functions as an amplifier. Figure 143 illustrates a circuit arrangement employing the grid-rectification principle.

When an incoming signal of modulated radio frequency is applied across the oscillatory circuit LC establishing an e.m.f. across the points GF , the plate current will be varied in accordance with the incoming modulated frequency; that is, the audio-frequency component will pass through the load impedance Z and the radio-frequency component through the reactance C_2 . In order that these conditions be fulfilled, the potential of the grid must be such that it will take current. Since

the audio plate-current variations depend almost entirely upon the audio-frequency voltages set up on the grid, it is necessary to build up a maximum voltage on the grid by inserting a small condenser C_1 in series with the grid. This condenser provides the necessary high impedance for the low-frequency modulated groups and a low impedance for the high-frequency signal variations.

A very important feature in connection with the condenser C_1 is that its reactance must be very large in comparison with the grid-circuit resistance. If this is not done, frequency distortion will result owing to the low grid impedance. The condenser, therefore, should have a capacity of between 0.0001 and 0.00025 mf. to offer a maximum IZ drop to the low frequencies and a minimum drop to the radio frequencies.

The purpose of the grid resistance or grid leak R_1 is to allow the accumulated charges of electrons on the grid, which gather during the positive halves of the signal voltage, to become dissipated, otherwise the grid would build up an excessive charge of electrons and become erratic in its behavior. Hence, it may be said that the grid-leak resistance prevents the grid from "floating" by allowing a continuous conduction of direct current during the intervals in which signal voltages are impressed upon the grid. It is apparent that the greater the incoming-signal voltage the larger will be the value of current flowing through R_1 ; consequently, its resistance value should be kept low in all cases where high-gain radio-frequency amplifiers are used preceding the detector tube. Furthermore, the larger the value of the signal voltage on the grid, the greater will be the negative grid bias due to the electron flow through the resistance from the filament to the grid when the grid is charged positively. Thus, when the grid bias increases, the grid-circuit resistance also increases, thereby raising the possibility of distortion at the higher audio frequencies. It is, therefore, essential that the grid-leak resistance should not exceed the value which will introduce serious audio-frequency distortion. Where higher values of resistances (above 3 megohms) are used, the grid-condenser value should be reduced to the order of 0.0001 mf. This greatly increases the sensitivity of the detector owing to the larger audio-frequency voltages which may be built up across a small capacity in view of its greater IZ drop.

137. Power Detection.—A power detector is essentially a plate-detection system in which the plate voltage and grid bias are so adjusted that a relatively larger audio-frequency power output may be obtained. A plate voltage of approximately 200 volts with a grid bias adjusted almost to the point of plate-current cut-off, about -20 volts, will give efficient results.

The important consideration for the proper operation of a plate power detector is that the grid voltage built up by the incoming signal must never be allowed to swing the grid positive. It is, therefore, necessary

that the signal voltage at 100 per cent modulation does not at any time exceed one-half the negative grid-bias voltage that is being used. The sensitivity of the detector, therefore, is greatly reduced, but the quality of reproduction with strong signals is evidently much superior to any other method of detection (except diode detection, Sec. 146).

The possibility of obtaining a fairly linear output across the external-load impedance with this method of detection is much better than with the grid-rectification system, particularly when receiving signals of 100 per cent modulation.

This condition may be clearly seen if we assume a modulated radio-frequency signal applied to the grid of a detector tube operating near the cut-off point. If this signal has its radio-frequency amplitude varied

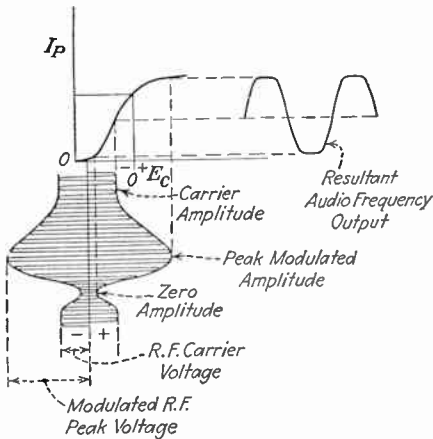


FIG. 144.—Distorted wave-shape characteristic.

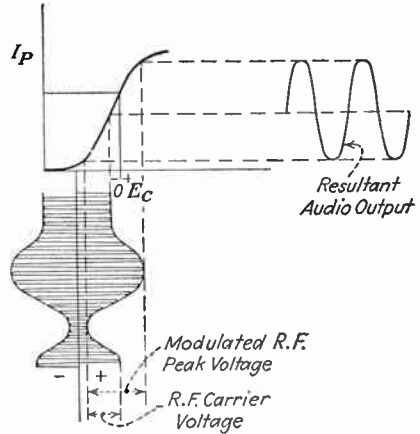


FIG. 145.—Symmetrical wave-shape characteristic.

to 100 per cent modulation, then the resultant audio-frequency output will depend on the I_p-E_c characteristic curve. If the curve should resemble a characteristic (curve) as illustrated in Fig. 144, the resultant audio frequency will not resemble a symmetrical reproduction of the signal envelope, but a box-type wave which will produce serious distortion. It is also evident that, if the C bias is operated slightly beyond cut-off, the lower portion of the audio output wave will also resemble a box-type characteristic. Great care must be taken, therefore, to use a tube which will operate along the linear portion of the characteristic curve so that the audio-frequency output will resemble a symmetrical wave shape of the modulated signal envelope as illustrated in Fig. 145. This condition may be obtained by preventing the modulated peak voltages from rising too high and extending into the curved portion areas. It is apparent, therefore, that low-percentage modulated radio-frequency waves, and waves having low radio-frequency peak voltages, may be expected to

provide the most symmetrical output. However, if the tube is properly designed so that the resultant I_p-E_c characteristic may have a longer straight portion, greater power output and good symmetry may be expected even with high-percentage modulated radio-frequency waves and high peak voltages.

138. Grid-leak Versus Power Detection.—Although it is possible to obtain linear detection by carefully adjusting the grid-condenser and grid-leak values, it is difficult thus to obtain good fidelity with the grid-leak detection method owing to the high value of R_p , particularly when functioning into reactive load impedances.

This form of detection is also undesirable in the reception of high-percentage modulated radio-frequency signals because the second-harmonic content of the modulated frequency will be greatly amplified owing to the square-law output of this detector. The second harmonic of the modulation frequency present in the plate circuit, therefore, will be equal to $\frac{1}{4} m^2$, where m is the modulation percentage of the received carrier frequency. In other words, considerable distortion will result when receiving high-percentage modulated signals because the relation between the modulated radio-frequency input and the audio-frequency output is not a linear reproduction of the signal-wave shape.

However, in view of the low plate voltage and resultant high R_p , the grid-leak type of detector is desirable where a high degree of sensitivity is required.

The power detector, on the other hand, although not nearly so sensitive as the grid-leak detector, will produce superior fidelity of the demodulated signal owing to its ability to obtain a better degree of linearity.

Furthermore, the power detector, owing to its lower sensitivity compared with the grid detector, permits less critical filtering of the succeeding audio-frequency circuits and also greatly reduces the possibility of microphonic amplification generally referred to as *singing*.

139. The Vacuum-tube Amplifier.—*a. The principles of the vacuum tube as an amplifier* of radio-frequency or audio-frequency currents are the same since both primarily amplify alternating-current voltages through the grid action of the tube affecting the space charge. The only difference is that in the audio-frequency amplifier the voltage of the input frequency ranges between approximately 16 and 15,000 cycles, while in the radio-frequency amplifier the input-voltage frequencies range from 20,000 to 60,000,000 cycles and higher.

There are certain important fundamentals, however, which are applicable to both audio- and radio-frequency amplifiers. Both systems are primarily energy-transferring devices in which the voltage or power of one circuit is transferred to a second circuit. In radio-frequency receiving circuits the energy transfer from one circuit to another is generally in terms of voltage amplification, since the vacuum tube is

primarily a voltage-amplifying device. The same holds true in audio-frequency receiving systems with the exception of the final amplifier stage where the energy transfer is generally one of power.

In order that the operation of the vacuum tube as either a voltage or a power amplifier may be clearly understood, the fundamental picture of the generator functioning into a load must again be applied.

Figure 80 illustrates the equivalent of a vacuum tube in a generator circuit. Units R_p and AC represent the vacuum tube as a generator unit. R_o and Z_o represent the external load or coupling unit into which the generator functions.

Although the grid is not shown, it is apparent that the generator symbol of the collector rings represents the operating condition of the tube which takes place during the period in which input voltages are applied to the grid. In other words, the generator, or plate circuit, may be visualized as being idle during the period in which no voltage impulses are received by the grid.

Now, since it has been shown that the maximum power from a generator can only be absorbed by a load if it is matched to the generator resistance or impedance, then the amount of energy that can be transferred from the plate circuit of a vacuum tube to its load must depend primarily upon the proper adjustment of the load impedance R_o or Z_o . Non-reactive loads containing neither inductance nor capacity are merely indicated R_o . Since most radio- and audio-frequency circuits contain reactive loads the symbol Z_o is generally used.

Hence it may readily be seen that the total amplifying property of a vacuum tube will be dependent upon several factors, namely,

1. The amount of signal voltage impressed upon the grid of the tube, E_g .
2. The amplification constant of the tube, μ .
3. The relationship of the load impedance to the plate resistance of the tube.

Perhaps one of the most important fundamental considerations in viewing the operation of the vacuum tube is to assume a signal e.m.f. E_g to be impressed upon the grid. This should be visualized as an e.m.f. μE_g acting in the plate circuit with which the tube resistance R_p and the load R_o or Z_o are in series.

This fundamental picture is illustrated in Fig. 85*b* in which the generator symbol represents the acting voltage of the tube μE_g into its load circuit R_p and L_1 . R_p is the plate resistance of the tube and L_1 the external load impedance.

Now since the voltage across the load impedance is primarily due to the input or signal voltages E_g , applied to the grid, the alternating current I_p in the plate circuit must be directly related to this voltage. Hence,

$$I_p = \frac{\mu \times E_g}{R_p + Z_o} \quad (117)$$

Therefore, the voltage or potential difference across the load impedance must be

$$E_o = \frac{\mu \times E_g \times Z_o}{R_p + Z_o} \quad (118)$$

The maximum voltage, therefore, will be built up across a load when the load impedance Z_o is infinite with respect to the tube plate resistance.

b. Complexities.—There are, however, many complexities which enter into the functions just described, particularly when the load impedance Z_o is reactive. This may be readily seen if we refer to the alternating-current engineering principle in which it is shown that the impedance if positively reactive (inductive) varies directly with the frequency and if negatively reactive (capacitive) varies inversely with the frequency. If, therefore, the external load of Fig. 80 is inductive, the ratio of the load impedance to the tube plate resistance will be a variable quantity and the output power or energy developed across the load impedance will also be dependent upon the frequency. In other words, high voltages will be developed across Z_o when the frequencies are high and low voltages when the frequencies are low.

Fortunately, however, if the load impedance is designed to have at least the same value as the plate resistance R_p at the lowest frequency that

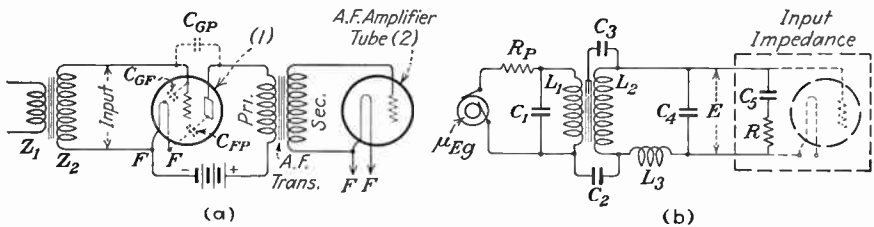


FIG. 146.—(a) Transformer-coupled audio amplifier; (b) equivalent circuit.

is to be amplified across Z_o , then the higher frequencies will be properly taken care of since they develop a larger voltage across the load due to the increase in the reactance X_L with the frequency. It has been found that a little over 85 per cent of the energy developed in the tube can be transferred or built up across a definite load impedance if the value of that load in ohms is approximately three times the plate resistance of the tube. Hence, if the load impedance is adjusted to give a maximum transfer at the lowest frequency to be received, a fairly flat output response may be expected over a band of frequencies.

If the load is inductive, as in the case of the audio-frequency transformer, the primary winding Z_o should contain a large value of inductance L to insure a maximum transfer of energy at the low audio frequencies below 100 cycles, thereby enabling a good audio-frequency response to be secured over the entire audible band, 30 to 10,000 cycles. There are, however, many other difficulties which arise in the audio-frequency transformer which limit the frequency response, particularly that of the distributed capacity in the windings.

*c. Transformer-coupled Audio Amplifier.*¹—Figure 146 illustrates a transformer-coupled audio-frequency amplifier circuit (a) and its electrical equivalent (b). C_1 and C_4 (b) represent the distributed capacity

¹ See also Sec. 220.

between the turns of the primary winding and the secondary winding, respectively. C_2 is the capacity existing between the two windings, C_3 is the capacity between the inner winding and the magnetic core, and L_3 represents the magnetic-leakage reactance. Although the distributed capacities may be kept extremely low by the proper design of the transformer, it is quite apparent that the higher range of audio frequencies will be by-passed because of the decrease in the circuit reactance at these frequencies.

Assuming that Fig. 146(a) and (b) represents two amplifier tubes coupled together by an audio-frequency transformer, then μE_g represents the voltage generated in the plate circuit of the first tube, R_p represents its plate resistance (also called *plate impedance*), and E is the voltage applied to the grid of the succeeding tube. Now, the secondary voltage E must be equal to the voltage across the primary winding multiplied by the ratio of the turns in the secondary. Hence, if the transformer has a ratio of 1:1 and the primary has a voltage of 1 volt built up across it, the transfer will represent an e.m.f. of 1 volt across the secondary. If the transformer ratio is 3:1 then the voltage generated in the secondary will be 3 volts.

Now, since it has been shown that the maximum power will be absorbed by the load when the impedance of the load is equal to that of the generator, it will be seen that the voltage built up across a reactive load impedance will vary with the frequency. Consequently, the voltage transfer will be a maximum when the load impedance matches the tube impedance at some one particular frequency. In other words, the impedance which is reflected into the primary circuit, so to speak, can be a match to the tube impedance at only one frequency. This is an ideal condition in a radio-frequency amplifying system in which only one frequency is to be amplified. Here the one frequency would produce the greatest voltage transfer, while frequencies above and below would have a decidedly falling-off characteristic.

For uniform amplification the voltage built up across the primary at each frequency should be very nearly the same over the entire audio-response range in order that a flat response may be obtained. Hence the design of the transformer must be such that there is no particular or sharply defined amplitude at one of the audio frequencies. This condition may be accomplished by resonating the transformer at some low audio frequency, say, 200 cycles, which would tend to flatten out the audio-frequency response characteristic.

The theoretical operation of a transformer circuit may be clearly seen if the circuit is modified to resemble a simple parallel-resonant circuit as illustrated in Fig. 147. Here L represents the effective primary and secondary inductance, and C , the total distributed capacities of the windings, including the grid-to-filament capacity.

If a low-frequency e.m.f. is applied across the impedance L , the circuit reactance will be due entirely to the inductance; but, if the frequency is increased and the inductive reactance also increases, the capacitive reactance due to C is decreased. At some point near the higher audio frequencies the capacitive and inductive reactances will be nearly equal, and parallel resonance results. If, therefore, the primary inductance of a transformer is made sufficiently high, and the secondary is not resonant at some low frequency, a fairly flat voltage-frequency response curve may be obtained.

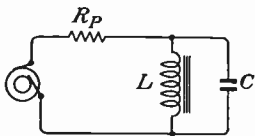


FIG. 147.—Modified equivalent circuit to Fig. 146.

If the inductance of the primary is too low, the resonant peak will be made much sharper because of the increase in the step-up ratio between the two windings. The resonant effect can be greatly reduced by using tubes having a low plate impedance since the effect is the same as lowering the number of turns in the primary load of the transformer.

Another important reason for maintaining a high primary impedance is the effect of the direct-current resistance of the primary winding absorbing a relatively greater amount of power than at the higher frequencies.

It is the loss caused by the direct-current resistance which generally results in a falling off of the response curve at the lower end of the frequency spectrum. The drop in the response curve at the higher end of the spectrum is due to the stray shunt capacities and leakage reactance acting as an attenuator.

A transformer may be designed to have a practically flat frequency characteristic over the entire audio-frequency band. This condition may be obtained if the resonant periods of the primary and secondary windings are slightly separated so as to produce a double-humped characteristic. If a resistance is then shunted across the secondary, the magnetic reactions upon the primary, due to the resistive load, will tend to flatten out the characteristic. With the careful design of the turns ratio, mutual inductance, and secondary load resistance to produce the proper magnetic reflections, the frequency-response curve may be made entirely flat from 40 to 10,000 cycles.

*d. Resistance-coupled Audio Amplifier.*¹—In this type of amplifier the theoretical operation of the circuit is somewhat different from that just explained for the transformer-coupled amplifier. In the resistance-coupled amplifier the voltage amplification or gain is that produced by the tube only and requires the use of high- μ tubes to effect the desired amplification. The theory of operation and general design considerations follow:

¹ See also Sec. 218.

When a signal is developed in the plate circuit of a detector tube in accordance with the general theory of detection, variation will take place in the plate circuit across the load or coupling resistance R_1 as illustrated in the schematic arrangement of Fig. 148(a). These variations will produce potential changes across R_1 in accordance with Ohm's law, $dE = dIR$. These variations, referred to as the *alternating component* in the plate circuit, are impressed across the input of the audio-frequency amplifier tube of an amplitude depending upon the product

$$E_1(\text{gain}) = \frac{R_1}{R_1 + R_p} \times \mu$$

It may be seen, therefore, by referring to the equivalent circuit in Fig. 148(b) that the voltage E_1 developed across R_1 must be applied without attenuation to the input or grid circuit of the amplifier tube GK if maximum efficiency is to be obtained. Consequently, if the coupling resistance R_1 is so chosen that its value is high compared with the tube plate resistance R_p , the gain will approach much nearer to the theoretical μ of the tube than if this is not done. However, since the variation in plate

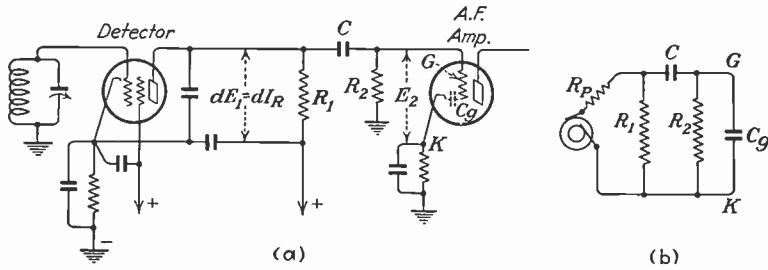


FIG. 148.—(a) Resistance-coupled audio amplifier; (b) equivalent circuit

current I_p will be dependent upon the ratio $I_p = \frac{\mu E_g}{R_p + R_1}$ it may readily be seen that the actual amplification of a tube, using a non-inductive load such as R_1 , can never be brought up to equal the theoretical value of μ . It is, therefore, simply a question of raising the load resistance as high as possible to approach the μ of the tube, but, as this will also require an increase in the plate-supply potential, some value which will not produce too large an IR drop must be chosen. The value of R_1 as generally used in the plate circuits of high- μ tubes of resistance-coupled amplifiers is usually between 10,000 and 1,000,000 ohms. The lower values of load resistance are used in stages where power output must be developed, and under these conditions a low- μ tube should be used. For practical purposes the load resistance R_1 , particularly when functioning in the output circuit of a detector tube of the variable- μ type, should be in the order of 250,000 ohms.

Now, in order that the maximum voltage developed across R_1 be applied to GK of the audio-frequency amplifier tube the values of C and R_2 must be so chosen that the voltage at E_2 will not only be of the same value as developed at E_1 at all frequencies (50 to 10,000 cycles), but must also produce a symmetrical reproduction of the signal-wave shape. Since the natural tendency for a grid condenser and grid leak is to introduce asymmetry, if some grid current flows, the possibility of distortion is apparent. The resistance R_2 must be sufficiently high to maintain a high input impedance between the points G and K , and, at the same time, must not be high enough to cause a piling up of electrons due to electron accumulation on the grid. The value of this resistance in general practice is between 100,000 and 500,000 ohms.

In the design of a resistance-coupled amplifier it is, therefore, important that the value of C be high compared to the amplifier-tube capacity C_g . This condition

may readily be seen by referring to Fig. 148(b), where the tube capacity and the resistance R_2 in series with C are effectively shunted across the plate load resistance R_1 . Hence the e.m.f. across C will not be in phase with that across the parallel branch GK since the latter contains resistance and capacity in parallel. Thus the voltage developed across G and K can only be a maximum, if the condenser C is made large so that its reactance is very small for the lowest frequency to be amplified. The value of C would not be critical for a large band of audio frequencies if it were not for the fact that the resistance R_2 is present. That is to say, this resistance forms a conducting load in the grid circuit which causes an IR drop and a subsequent drop in the voltage developed across GK . Hence, in order that a maximum voltage amplification may take place over a wide band of audio frequencies, say from 50 to 10,000 cycles, the value of C together with its associated resistances must be of the proper value. To accomplish this, a formula has been developed¹ which will enable the proper selection of these values:

$$\text{Voltage ratio, or } K_{max} = \frac{\frac{R_1 \times R_2}{R_1 + R_2}}{\frac{R_1 \times R_2}{R_1 + R_2} + R_p} \quad (119)$$

On the basis that the voltage-transfer variation does not vary in excess of 10 per cent over the entire audio-frequency range, it will be essential to choose a percentage of the voltage ratio at the lowest frequency to be amplified (50 cycles). This is accomplished by taking the square root of the value K . This value should be between 90 and 97 per cent if a uniform response over the entire range of frequencies is to be obtained. If two stages of amplification are to be used, each stage must be uniform within these limits or, in other words, near the square root of 90 per cent which is 94.9 per cent. Hence the value of the coupling condenser may be computed by the ratio

$$C = \frac{R_1 + R_p}{6.28 \times f[R_2(R_1 + R_p) + (R_1 \times R_p)] \sqrt{\frac{1}{K^2} - 1}} \quad (120)$$

where K for a single stage is $\sqrt{0.90}$ or 0.949 (approximately 95 per cent).

Summarizing, it may readily be seen that, if a large value of R_2 is used, the coupling condenser C may be small. This is a desirable condition since smaller values of coupling capacities reduce the possibility of audio-frequency oscillation. The values generally used range between 0.01 and 0.1 mf . If higher values are chosen, suitable plate and grid filtration must be employed to prevent the possibility of self-oscillation.

If the fidelity range of a resistance-coupled amplifier is to cover a wide band of frequencies up to and above 10,000 cycles, it will be necessary to overcome the attenuation at these frequencies, owing to the by-passing effect of the input tube capacity C_o , by inserting a parallel-resonant or equalizing circuit across GK tuned to approximately 10,000 cycles.

e. Impedance-coupled Amplifier.—In the impedance-coupled audio-frequency amplifier the circuit arrangements as shown in Fig. 149(a) are

¹ HARRIS, SYLVAN, *Proc. I.R.E.*, December, 1926.

identical with those of the resistance-coupled amplifier with the exception that the plate load is reactive, Z_1 , instead of resistive, R_1 [Fig. 148(a)]. A grid-leak resistance R_2 may be used as in the resistance-coupled amplifier, or a reactance leak Z_2 may be employed.

The theoretical operation of this amplifier in the plate-load circuit is identical with that of the primary winding of the audio-frequency transformer previously explained, with the exception that the load functions into a series-resonant circuit as illustrated in Fig. 149(b). It is seen that if an impedance Z_2 is used, the combination of C and Z_2 in series will be resonant at some audio frequency. If C and Z_2 are so chosen that they provide series resonance at say 100 cycles, a maximum voltage will be developed across G and K and in the vicinity of these frequencies. This will increase the response at the low end of the frequency spectrum.

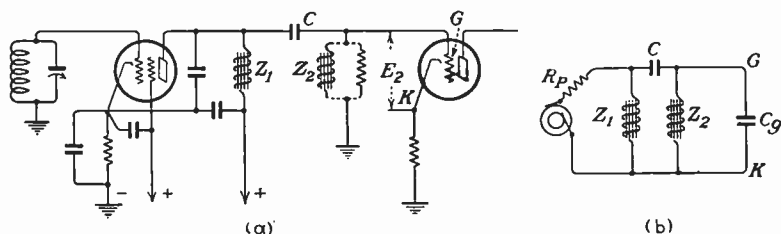


FIG. 149.—(a) Impedance-coupled audio amplifier; (b) equivalent circuit.

The advantage of this type of amplifier is due to the low IR drop in the plate-load winding Z_1 as compared with the resistance-coupled type, and also to the high voltages that may be developed across Z_1 due to its reactive function.

Alternating-current engineering principles may readily be applied to Figs. 148(b) and 149(b) to determine the voltages developed across the component reactances and impedances.

140. Amplifier Distortion.—Amplifier distortion may be due to any one or all of the following conditions.

1. Excessive grid swing.
2. Improper grid bias.
3. Improper plate voltage.
4. Generation of undesirable harmonics.
5. Core saturation (audio-frequency transformer).
6. Improper load-impedance matching.
7. Improper reflected impedance from secondary circuits.
8. Improper filtering.
9. Self-oscillation.
10. Improper balancing in push-pull circuit.

Any one of these conditions may cause an alternation of the plate curve characteristic to resemble an unsymmetrical wave shape. Let us take, for example, the result when an insufficient grid bias is applied to

the tube as shown in the plate-current characteristic illustrated in Fig. 150(a). If the bias is excessive, the plate-current shape will take the form of diagram (b). If the bias is correct, but the signal voltages are excessive, the plate-current shape will resemble Fig. 144. If the plate and grid voltages are correct, but the load impedance is too low, the plate-current shape will resemble diagram (c).

All of these conditions will produce distortion and the generation of harmonic frequencies.

The total distortion in any amplifier is practically entirely dependent upon the magnitude of the second and third harmonic-frequency components. Although it is possible to reduce these components by operating the tube into a load impedance relatively higher than that of the R_p of the tube itself as with the power triode, or about one-seventh with the

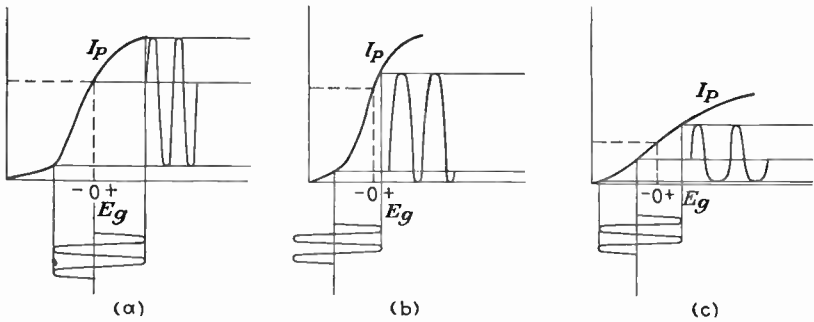


FIG. 150.—Plate-distortion characteristics.

power pentode, to obtain a fairly linear output, the harmonics can never actually be completely eliminated. Hence the only remedy would be to maintain the total harmonic-frequency contents of the amplifier as low as possible to obtain negligible distortion.

If the combined second and third harmonic-frequency energy components do not exceed more than 5 per cent of the fundamental frequency, a minimum degree of distortion may be expected. The total percentage of harmonic distortion may be obtained by taking the vectorial addition of the second and third harmonics:

$$\text{Distortion} = \sqrt{H_2^2 + H_3^2}$$

The harmonics H_2 and H_3 are obtained by taking the ratios of the current components due to the normal-operating plate current and the changes in current due to the harmonics, thus

$$\text{Distortion} = \left[\frac{1}{2} \times \frac{(I_{max} + I_{min}) - I_{normal}}{I_{max} - I_{min}} \times 100 \right] \text{ per cent} \quad (114)$$

All higher harmonics are generally neglected since their amplitudes are too small to produce noticeable distortion.

It is possible practically to determine if excessive distortion is present due to harmonics in the power stages of an audio-frequency amplifier by inserting a milliammeter in the plate circuit. Since the generation of harmonics develops an asymmetrical condition in the plate circuit resembling rectification, the milliammeter will rise or decrease above its normal value while signal voltages are being received. Hence, if the plate milliammeter fluctuates above or below its normal value during reception, it is evident that rectification is taking place and distortion is present. Serious harmonic distortion will produce heavy fluctuation.¹

141. Power Amplification.—It has been explained that the vacuum tube functioning as a voltage amplifier delivers an output voltage dependent upon the product of the input voltage times the amplification factor of the tube. For example, if a potential of 1 volt is built up across the grid-filament circuit of a tube having an amplification factor of 10, the output voltage would be $E_o = \mu \times E_g = 10 \times 1 = 10$ volts.

Hence, if the input voltage is increased, the output voltage will increase in proportion. In the power amplifier, however, the output power is proportional to the square of the input voltage, consequently greater input voltages are required for effective operation. Now, if it is desired to convert the voltage into power in the output circuit of a vacuum tube, energy must be dissipated in the plate circuit by introducing a load which will be capable of absorbing power. If this load is too highly resistive or reactive, very little power can be absorbed and consequently very little energy can be transferred. Hence, if power is to be developed in the output circuit of a vacuum tube the load resistance or impedance must not be too high. Thus, if a low-resistance load circuit is required to absorb power, a vacuum tube of a low plate resistance must be used, as the maximum power will be transferred to the load impedance only when its ohmic resistance is equal to that of the tube impedance.

The maximum power output in watts developed across the load of a tube may thus readily be determined by the ratio

$$P_o = \frac{\mu^2 \times E_g^2 \times Z_o}{(R_p + Z_o)^2} \quad \text{watts} \quad (121)$$

where Z_o represents the output load resistance or impedance and E_g the input grid voltage developed across the grid filament of the tube.

Let us take for example a power tube such as the W.E. 205E which has an amplification constant of 7 at 300 volts and a plate impedance, R_p , of 4,000 ohms. If an average input voltage of say 20 volts is applied to the grid, and the load impedance is made to match the tube impedance, the output power into the load impedance will be

$$P_o = \frac{49 \times 400 \times 4,000}{(4,000 + 4,000)^2} = 1.2 \text{ watts (approx.)}$$

¹ A method for measuring harmonic distortion has been developed by C. G. SMITH *Proc. I.R.E.*, vol. 18, p. 178, January, 1930.

If the load-impedance values in both the numerator and the denominator in the above example are changed to higher or lower values than the tube impedance, it may be clearly seen that this impedance mismatch will develop considerably less power in the load than under the matched impedance conditions. Thus it may once more be seen how the matching of tube impedances is essential where a maximum degree of power transfer is desired.

In high-quality audio-frequency amplifying systems, however, the matching of load and tube impedances may not be desirable because of the increased possibility of distortion, particularly if the load impedance is too low. This is due to the fact that, when the load impedance is too low, the straight portion of the I_p - E_c characteristic curve may be considerably shortened. Consequently, if fairly large input voltages are applied to the grid under these conditions the resultant plate-current variations may resemble a box type or an asymmetrical signal-wave shape as illustrated in Fig. 144. This will result in the production of a second-harmonic or double-frequency component supplemental to the original signal input. Hence, if this second-harmonic content is too large, serious distortion of the original signal-wave shape will occur. To prevent this, the load-impedance and operating characteristics must be carefully analyzed to obtain a *maximum undistorted* power output.

Now, if the load impedance is made greater than the tube impedance the output power developed across Z_o will also be considerably less as has been shown, but, on the other hand, it will tend to increase the length of the straight portion of the I_p - E_c characteristic curve. Hence, if the plate impedance is approximately twice the impedance of the tube, the maximum degree of undistorted power with a minimum second-harmonic content will be obtained. This condition will, however, not be entirely fulfilled unless the signal-input grid and bias voltages are also increased.

Tubes operating into matched impedances provide the greatest amount of power output, whereas tubes operating into higher load impedances develop less power but provide a greater linear characteristic. Hence, tubes operating into loads of twice their internal impedances provide a greater possibility of obtaining a symmetrical reproduction of the input signal voltage by the extension of the straight portion of the plate characteristic curve and thereby minimizing the degree of distortion of the incoming signal.

The maximum possible undistorted power output, when the load impedance is twice the tube impedance, may be obtained from the ratio

$$P_{max} \text{ (undistorted)} = \frac{\mu^2 \times E_g^2}{9 \times R_p} \text{ watts} \quad (122)$$

Perhaps two of the most important factors in the design of power-amplifier systems to prevent distortion is, first, to design the load imped-

ance so that the impedance at all times will remain higher than the tube impedance, particularly when the lowest audio frequency is being impressed across it. This will require a tube having a low plate resistance. Second, to maintain a value of grid bias which will at all times be greater than the maximum peak voltage of the signal swing. For example, if an incoming signal of 20 volts is applied to the grid filament of the tube, the negative grid-bias potential should not be less than -20 volts and preferably more.

If the signal swing exceeds the bias potential, a grid current will flow during the incoming positive peaks, and a *resultant* load is placed upon the plate circuit. This effect functions upon the $I_p - E_c$ characteristic curve the same as decreasing the negative bias, because, when the grid draws current and an IZ drop takes place across the filament-grid circuit, it reduces the voltage built up across these two points. As soon as this

G_1 = Control Grid; G_2 = Screen Grid; G_3 = Suppressor Grid

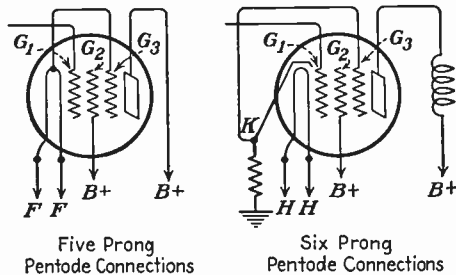


FIG. 151.—Pentode-tube connections.

takes place, the operating point on the curve is moved up toward the top portion of the characteristic, so that the incoming signal produces an asymmetrical plate-current¹ effect owing to the loss in the linearity of the curve. Thus it may be seen that if the operating point is moved upward on the $I_p - E_c$ curve the impedance of the plate circuit will be lowered considerably.

142. The Pentode Tube.—Figure 151 illustrates two arrangements of the pentode variety of screen-grid tubes. In addition to the screen grid previously explained, another grid called a suppressor is placed close to the plate element in the tube. This grid may be connected directly to the filament or the cathode of the tube as illustrated in the figure, thereby placing it at ground or zero potential. In some types of tube this connection is made internally, to either the filament or the cathode, depending upon the type used. Hence no difference of potential exists between this grid and the filament, since both are at the same electrical level or potential. Its function, however, is extremely important owing

¹ The term *asymmetrical* or *non-symmetrical* referred to in vacuum-tube circuits is used to express the lack of equal proportion between the positive and the negative halves of a voltage or current cycle.

to its effect upon the negative field produced in the immediate vicinity of the plate.

When electrons are bombarding the plate in tremendous numbers, in tetrode and pentode tubes, the impact of the electrons against the plate causes the plate to emit electrons. This effect is called *secondary emission*. These electrons produced at the plate may be collected by the screen-grid, causing an increased flow of screen-grid current and a consequent reduction or dip in the plate characteristic. This dip will produce serious plate distortion and resultant harmonic generation. If, however, a grid is located very near to the plate and at ground potential, the emission of secondary electrons from the plate will be repelled. This prevents the secondary electrons from reaching the screen-grid and reduces the possibility of plate distortion. In addition the repelling action of the suppresser grid permits a relatively larger development of power in the plate circuit without distortion.

The pentode tube as a power amplifier in receiving circuits, therefore, has some advantages over the triode type of amplifier, particularly in that the tube has the ability to develop slightly more power and requires only about one-third of the signal voltage for a given amount of power.

Hence, since the amplifying ability of the tube is relatively greater for a given impressed signal voltage, it is possible to eliminate one of the preceding voltage-amplifier stages and thereby reduce the number of tubes and the possible distortion inherent in all amplifiers.

143. The Power Pentode Load.—It has been previously explained that the output load of the vacuum tube should at least be equal to the plate impedance (R_p) of the tube, if a maximum power is to be developed in the load circuit. It was also shown that the output load of an audio-frequency power-amplifier triode should be twice the plate resistance of the tube, if fidelity is to be obtained at the low audio-frequency spectrum. This is mainly due to the low plate resistance of the power triodes, between 1,500 and 3,000 ohms.

The power pentodes, on the other hand, have a considerably higher plate resistance, approximately between 10,000 to 60,000 ohms. If, therefore, the load-impedance values were based upon triode conditions, considerable distortion would result owing to the increased amplification of the harmonic content of the signal frequency. Hence, in order to reduce the power developed by the harmonics in the output-load circuit, a much lower value of load resistance relative to the plate resistance is used. The optimum value of the load resistance in power pentodes is generally in the vicinity of 7000 ohms.

The maximum undistorted power output of a pentode may be calculated from the formula

$$P_{\text{undistorted}} = 0.33 \times E_p \times I_p = 2.5 \quad \text{watts (approx.)}$$

The constant 0.33 or one-third is derived from the plate efficiency of a pentode when operating into a low load impedance. Under these conditions the plate efficiency is approximately $33\frac{1}{3}$ per cent.

For example, where the output-load resistance for a triode power amplifier having an R_p of 2,000 ohms must be at least 4,000 ohms, the output-load resistance of a power pentode with a plate resistance of 60,000 ohms would be approximately 7,000 ohms. These conditions must be thoroughly fulfilled if good quality and power are to be obtained. The value of the output-load resistance of a pentode tube may be calculated approximately from the formula

$$R_o = \frac{E_p}{I_p}$$

where E_p = the operating voltage and I_p = the operating plate current; it being assumed that the bias is correctly adjusted for these conditions.

For example, the -47 pentode tube requires a plate voltage E_p of 250 volts, a bias voltage E_c of -16 volts, and a screen voltage E_{so} of 250 volts. Under these conditions a plate current of 30 ma. (0.030 amp.) will flow. Hence the approximate value of the load resistance for this tube will be

$$R_o = \frac{E_p}{I_p} = 8,000 \text{ ohms (approx.)}$$

144. Push-pull Amplification.—This form of amplification is also known as *balanced amplification* and refers to the distribution of voltages and currents in vacuum-tube grid and plate circuits. This method of amplification is generally used in radio- and audio-frequency power-output circuits to develop a symmetrical or balanced wave shape in the load circuits together with an increase of power output. Tubes connected in this manner also serve to reduce the generation of the undesirable second-harmonic frequency which is normally developed in single-tube output circuits due to asymmetry, resulting from improper circuit adjustments and characteristics.

It has previously been shown that the output circuit of any amplifier must reproduce symmetrically the grid voltages impressed upon the grid so that the output-circuit variations conform with that of the input-signal shape.

We have seen that this plate symmetry can only be approached provided that the grid bias is so adjusted that the tube will operate at the center of the straight portion of the I_p - E_c characteristic curve. This is no assurance, however, that the output-wave shape will be a symmetrical reproduction of the signal on the grid since an excessive grid swing may drive the plate currents into the upper and lower curved portion of the I_p - E_c characteristic and produce a square-topped type of wave as previously explained. Furthermore, if the load impedance is too low,

distortion will also result owing to the asymmetrical variations developed across the impedance. For example, let curve (a) in Fig. 152 represent an incoming wave developed across GK of a simple amplifier circuit, diagram (d). If the grid bias is adjusted at the straight portion of the curve, and no grid current is flowing, the plate-current curve will be that illustrated by the heavy curve in (b). In other words, the increase in plate current, when the grid is positive owing to the signal, will be relatively greater in amplitude than the decrease in plate current when the grid is negative owing to the signal. Consequently, the signal will not be represented by a symmetrical variation in the plate circuit and distortion will result.

If the input transformer is so designed that the grid circuit is continually drawing some current the output-wave shape may be more

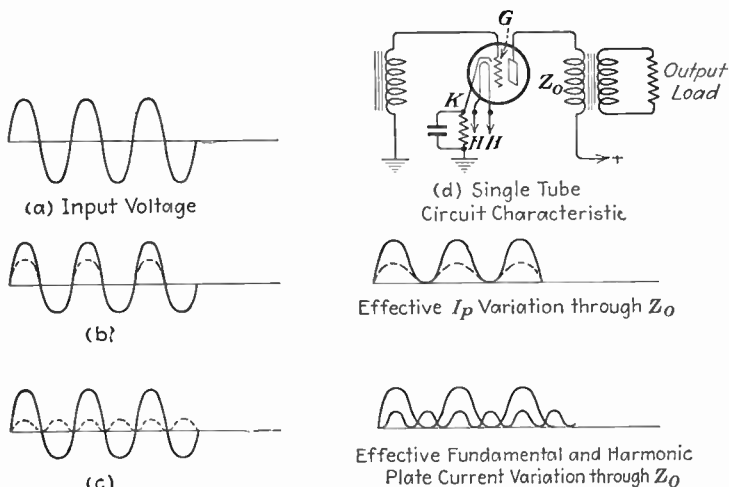


FIG. 152.—Single-tube circuit and its fundamental and harmonic components.

symmetrical as illustrated in (b) by the dotted curve. This is due to the fact that, when the grid current flows during the positive half of the input cycle, an IR drop takes place through the resistance of the secondary winding of the input transformer. Hence, the amplitude of the plate current during the positive peak will be reduced thereby producing a more symmetrical output in the plate circuit. However, since the grid circuit absorbs power under this condition, less amplification will be available in the output-load circuit Z_o in Fig. 152(d). It has been previously shown, however, that if distortion is due to a low-plate load impedance, it may be practically eliminated by raising Z_o to a value at least equal to that of the tube resistance, R_p . With a single tube, therefore, it is difficult to produce a symmetrical plate-current output unless the signal e.m.f.s. are kept very low and the plate-load resistance high with respect to the tube resistance. Hence, the ever present possibility of

asymmetrical plate-current variations will produce a wave form in the plate circuit equal to twice that of the fundamental frequency. This is called the *second-harmonic frequency* and is illustrated by the dotted curve in (c). Obviously, the combining of the fundamental and harmonic frequencies through Z_o will produce distortion. This is one of the most

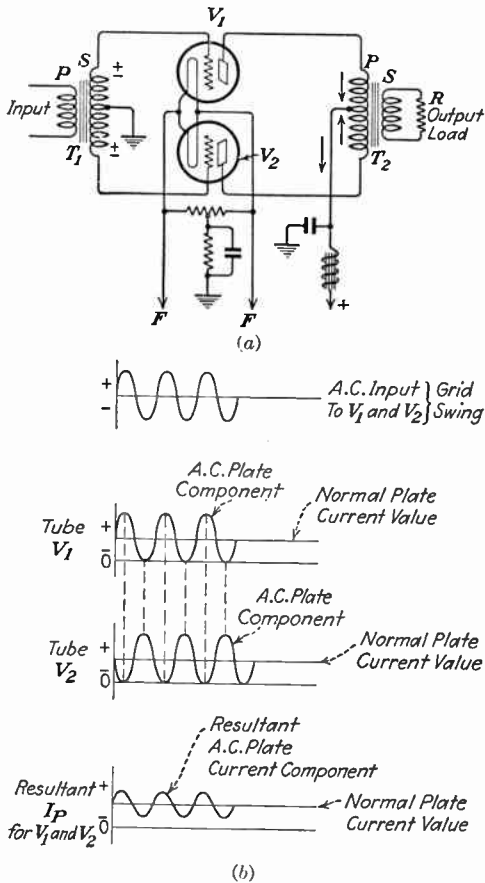


FIG. 153.—Push-pull amplifier: (a) circuit; (b) operating characteristics.

prevalent causes of distortion in single-tube amplifiers, particularly if large grid swings are applied.

In the push-pull or balanced amplifier it is possible to balance out this second-harmonic frequency by the output transformer windings so that the fundamental frequency alone will be present in the output secondary winding or load circuit. How this is accomplished will be explained in the following analysis of a balanced amplifier.

Theory of the Push-pull Amplifier.—Figure 153 (a) and (b) illustrates the conventional circuit arrangement of a push-pull amplifier and wave characteristics as used in an audio-frequency power-amplifier system.

If an audio-frequency e.m.f. is developed across the primary winding of the input transformer T_1 , a voltage will be induced across the secondary winding of the transformer in accordance with the fundamental laws of induction previously explained. Consequently, during each half of the cycle the two grids are of opposite polarities; but, since this polarity is constantly changing, the two grids are continually alternating in polarity with respect to each other. Thus when the grid of V_1 is positive the grid of V_2 will be negative and vice versa.

Now, let us assume an incoming audio-frequency signal of 1,000 cycles is being developed across the secondary winding S of the input transformer T_1 . This signal will apply to the grids of both tubes an e.m.f. of equal magnitude but with a phase difference of 180 degrees. That is to say, the voltages developed on both of the grids will be 180 degrees out of phase with each other. Now, since the grid of one tube will be positive and that of the other tube negative during each cycle, the alternating plate voltage in the primary winding of the output transformer will also be 180 degrees out of phase with each other. Consequently, the resultant plate-current component in the two halves of the primary winding will also be 180 degrees out of phase with each other. However, the magnitude of the current flow in each half will be of a different value, since the plate-current flow in the primary half of V_1 , will be higher than that of V_2 , if the grid voltage at this instant is positive on V_1 . Then, as the cycle reverses to make the grid of V_2 positive and of V_1 negative, the plate-current magnitudes will be such that the amplitude in the plate half of V_2 , will be greater than that of V_1 . Now, since the currents in the two halves are in opposite directions, it would seem that the magnetic fields created by them would cancel each other. This condition would prevail if the currents in each half of the winding were of equal magnitude, but, owing to the fact that the current flow in each half is dependent upon the polarity of the grid potential of each tube, it is apparent that the plate current in V_1 will be greater than that in V_2 during the instant when the grid of the former is positive and that of the latter negative. Hence, the effective magnetic field created around the primary winding of T_2 will be the difference between the current flowing in the primary winding of V_1 and V_2 . In other words, a signal voltage of 1,000 cycles will be developed across the primary winding of the output transformer.

Now, if the secondary winding of the output transformer is open so that no current and magnetic field are produced, no reaction will take place upon the primary winding of T_2 . Under this condition the second harmonic-frequency current component in both halves of the primary winding will be in opposite directions causing the magnetic fields to oppose and cancel each other. The fundamental frequency, on the other hand, will be increased since the magnetic fields due to it are in the same direction. These conditions are due to the fact that the currents in the plate-circuit halves, due to the harmonic frequency, are in phase, while those due to the fundamental frequency are 180 degrees out of phase with each other. Now, if the secondary winding is closed by a load resistance R to represent a power-operated device such as a loud-speaker or telephone line, the voltages developed across the winding will cause a current of an alternating character to circulate through it at a frequency depending upon the arrangement of the secondary windings.

For example, if the input frequency of 1,000 cycles is producing plate variations in the two halves of the primary winding, the e.m.f. developed across the secondary winding will depend upon the manner in which the secondary is wound. That is to say, if the secondary winding forms a continuous coil as illustrated in Fig. 154(a), the e.m.f. developed across it will be proportional to the difference between the plate currents in the primary halves. This is due to the fact that the relationship of the secondary winding to the two halves of the primary windings is such that the two halves of the secondary will have e.m.fs. developed in opposition to each other, since the current in the primary halves are 180 degrees out of phase. Thus the e.m.f. and

resulting current flowing in the load circuit will be the proportionate difference between the two e.m.fs. of the two plate-current components. During each half cycle, therefore, the e.m.f. developed across the secondary will resemble amplitudes of an alternating character, such as illustrated in Fig. 155. Thus the original signal frequency of 1,000 cycles will alone be developed across the secondary winding, since the harmonic has been previously canceled out in the two halves of the primary winding. Hence, in a balanced or push-pull arrangement, the output e.m.f. will approach the ideally symmetrical output even though each half of the cycle is asymmetrical in character as illustrated in Fig. 155.

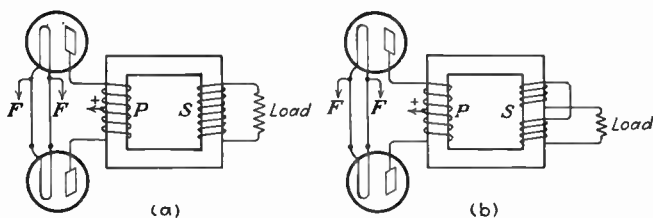


FIG. 154.—Push-pull circuits: (a) output circuit for second-harmonic elimination; (b) frequency-doubling circuit.

Now, if the secondary winding is arranged, as in Fig. 154(b), so that the two e.m.fs., due to the two plate-current variations, are additively developed across the secondary, an e.m.f. of twice that of the 1,000-cycle fundamental frequency will be developed. In other words, the latter arrangement will cause the magnetic fields of the 1,000-cycle frequency to cancel owing to their phase opposition; while the magnetic fields due to the harmonic frequency will be additive because they are in phase. The balanced amplifier is then functioning as a frequency-doubling device.

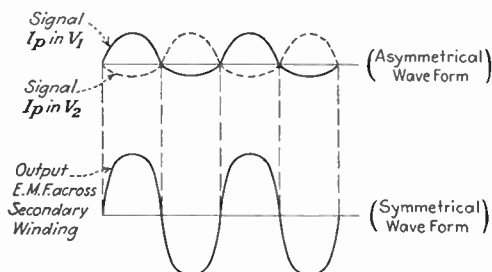


FIG. 155.—Symmetrical wave form produced by a push-pull amplifier.

In addition to its ability to balance out the inherently generated second-harmonic frequency and produce a symmetrical signal output, the push-pull amplifier also possesses several other advantageous features. These are summarized as follows:

1. The power output of a push-pull amplifier is very nearly equal to that of three tubes connected in parallel. This condition, however, is only true if the load impedance is properly designed to match the effective plate resistance R_p of the two tubes.

2. Owing to the fact that the direct-current plate components are always in opposite directions, the resulting magnetic fields cancel. In audio-frequency output transformers this is a highly desirable condition because it prevents the magnetic saturation of the iron core by the normal d.-c. plate-current components.

3. The possibility of a hum frequency due to the ripples developed in the power-supply system is practically eliminated, since these ripples will be in phase in the respec-

tive halves of the primary winding of the output transformer and will therefore cancel each other.

145. Grid and Plate Filtration.—Another important consideration in the design of all types of amplifying systems, including both radio- and audio-frequency amplifiers is the proper isolation between coupled stages to prevent interstage coupling and reactions. This is particularly important in high-gain amplifiers where the possibility of common coupling (common impedance) is prevalent owing to the use of a single power supply for all stages. Common coupling may produce serious distortion in amplifiers owing to the circuit reactions causing variations of grid voltages at different frequencies, thereby distorting the signal-wave shape. In many cases the reactions may be great enough to produce whistling and howling, particularly in high-gain radio- and audio-frequency amplifiers.

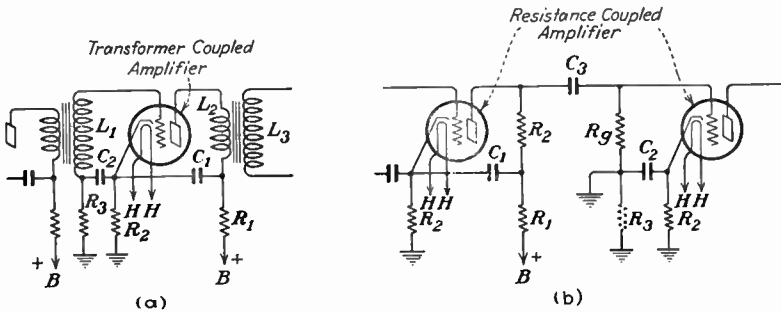


FIG. 156.—Grid- and plate-filtering circuits.

Figure 156(a) and (b) illustrates a typical audio-frequency amplifier using grid- and plate-filtering systems. The plate-filter system consists of the resistance R_1 and the condenser C_1 ; and the grid filter, diagram (a), of the resistance R_3 and the condenser C_2 .

The theoretical function of the filter circuit may be more readily understood by referring to the fundamental circuit in Fig. 157. Here the alternating-current generator in series with the plate supply represents the irregularities (ripple) in the plate circuit, caused by the power supply. If at 120 cycles the condenser C has a reactance of 1,000 ohms, the impedance of the combination will be about one-third that of the resistor alone, and the resultant current drop across it will be correspondingly less. As no direct-current flows through the 100,000-ohm resistor in series with the grid, there is no direct-current voltage drop across it. Thus, it has no effect on the bias. It will be noted that the high resistance and the condenser form a voltage-dividing circuit for the alternating current appearing across the biasing resistor. Only those portions of the alternating-current and audio-frequency voltages appearing across the condenser C will be impressed on the grid of the tube.

As the reactance of the condenser at 120 cycles is 1,000 ohms, compared to the 100,000 ohms resistance, roughly only one one-thousandth of the alternating-current or ripple voltage appearing across the biasing resistor will reach the grid.

In Fig. 158 is shown the adaptation of this circuit as usually used for the *input* tube of a high-gain system. This point is, of course, the most critical in regard to hum. The first filter C_1, R_1 reduces the alternating-current to one one-thousandth of its original value. This e.m.f., which appears across condenser C_1 , is impressed on the second and similar filter circuit C_2, R_2 and is further reduced by the same proportion; so the final reduction is 1,000,000:1. As no grid current is allowed to flow, the bias voltage is unaffected by the high resistors.

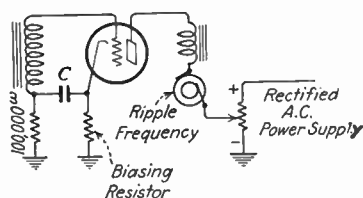


FIG. 157.—Illustration of principle of grid-filtering circuit.

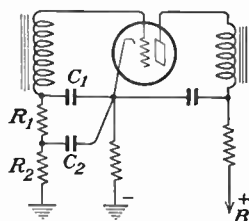


FIG. 158.—Grid-filtering circuit for high-gain audio-input circuits.

The complete stabilization that may be obtained in amplifiers using these methods has resulted in its wide adoption among amplifier-designing engineers.

Note: For further discussion on plate filters see Sec. 91 on Radio-frequency Chokes.

146. The Diode Detector.—Figure 159 illustrates the circuit of a simple two-element rectifying system using the cathode-heater type of tube in the diode circuit. The cathode and the plate are tied together to form one element, and both are connected to ground potential. No plate voltage is used in order that the rectified-signal group may be entirely dependent upon the signal-wave shape itself.

Let us assume that an incoming-signal frequency of a modulated character is amplified through a number of radio-frequency amplifier stages and applied across the tuned circuit XY when this circuit is properly adjusted to resonance with the incoming frequency. Since the incoming signal is of a damped character, its amplitude will vary in accordance with the modulated frequency as illustrated in Fig. 160(a). However, owing to the rectifying action of the diode, the wave shape across C_2R_1 and R_2 will be of an asymmetrical characteristic, such as illustrated in Fig. 160(b). The reason for this asymmetry is as follows.

When the positive halves of the incoming cycle are applied to the grid, electrons will be attracted to it and will flow in the direction of the arrows through L_1, R_1 , and R_2 , returning to the cathode and completing

the circuit. The flow of electrons through the resistance R_2 builds up an e.m.f. across it. As the negative half of the cycle approaches, no electrons will be attracted to the grid, and consequently no current will flow

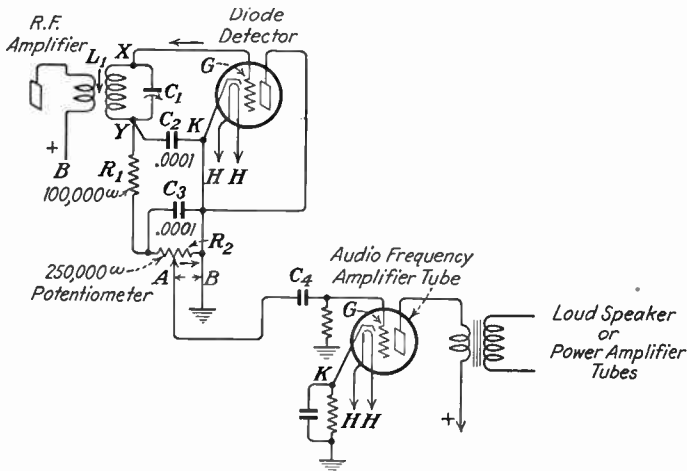


Fig. 159.—Diode-detector and audio-amplifier circuits.

through the circuit $L_1R_1R_2$. As the cycle reverses, making the grid positive, electrons will flow through R_2 , again establishing an e.m.f. across it. However, since the amplitude of the second positive cycle is

of a different value (lower in this case) the e.m.f. built up across R_2 will be relatively lower. Hence, if the sequence of the positive wave peaks varies at an audible rate, the resultant e.m.f. across the resistance R_2 will represent an audio-frequency variation in accordance with the group frequency of the modulated radio-frequency signal (Fig. 160(c)).

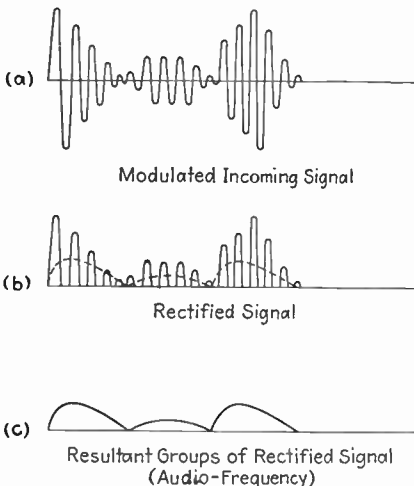


FIG. 160.—Incoming and rectified wave characteristics.

The resistance R_1 and the condenser C_3 constitute a conventional grid-filtering system. This combination permits a maximum radio-frequency signal voltage to be applied across the grid-cathode elements through the by-pass condenser C_2 .

The condenser C_2 , which is also a radio-frequency by-pass, builds up an e.m.f. across itself during the positive halves of the signal cycle, and discharges through R_2 in pulses, depending upon the group frequency of

the modulated signal. It is the accumulated discharge of this condenser, which in effect represents the audible or group voltage built up across R_2 . If C_2 were not provided, practically no effective group voltages could be built up across R_2 , and the resultant audio-frequency signal would be negligible.

Generally the value of this condenser ranges between 0.0001 and 0.002 mfd. If the value should exceed 0.002 mfd., insufficient radio-frequency voltages would be developed across it.

Thus it may be seen that the effectiveness of a rectifier circuit will depend almost entirely upon the radio-frequency by-pass condenser C_2 , if a high value of radio-frequency voltage is to be built up, in order that the pulse or group frequency [curve (c) Fig. 160] may be of sufficient amplitude.

The rectified low-frequency voltages may then be suitably amplified by feeding them to an audio-frequency voltage-amplifying system. These voltages are taken off at the points A , B of the potentiometer R_2 , and are regulated in accordance with the volume required.

The audio frequencies are fed to the first amplifying tube through the medium of the coupling condenser C_4 . Hence, any audio voltages developed at the points A , B will be applied directly across the GK elements of the amplifier tube.

The diode system of detection unquestionably provides the finest form of rectification with negligible distortion. The disadvantage is primarily in the low amplification secured, as compared with the triode type of rectifier. If, however, sufficient radio-frequency amplification is provided ahead of the diode detector, large signal voltages may be developed, which will insure a desirable audio-frequency output.

147. The Automatic Volume Control.—Figure 161 illustrates a vacuum-tube circuit in which the incoming signal volume is so controlled that the overall gain or signal level is practically constant. This eliminates excessive blasting and overloading at the high incoming peak voltages and serves to improve the fidelity of reception. The operation of the tube in producing this function is called the *automatic-volume-control* or *AVC* circuit.

a. The theoretical operation of all *AVC* circuits is based upon the biasing control of the radio-frequency amplifier grids. That is to say, when incoming signal voltages are amplified through a number of radio-frequency stages the normal negative grid bias is reinforced by an additional negative bias which increases in direct proportion with the increase of the signal voltage. As the signal voltage increases, the grid bias is automatically increased to the point in which the signals are actually "throttled." By the careful adjustments of the *AVC* circuit constants all excessive signal peaks may be kept down to a desirable level and the possibility of overloading and distortion thereby greatly minimized.

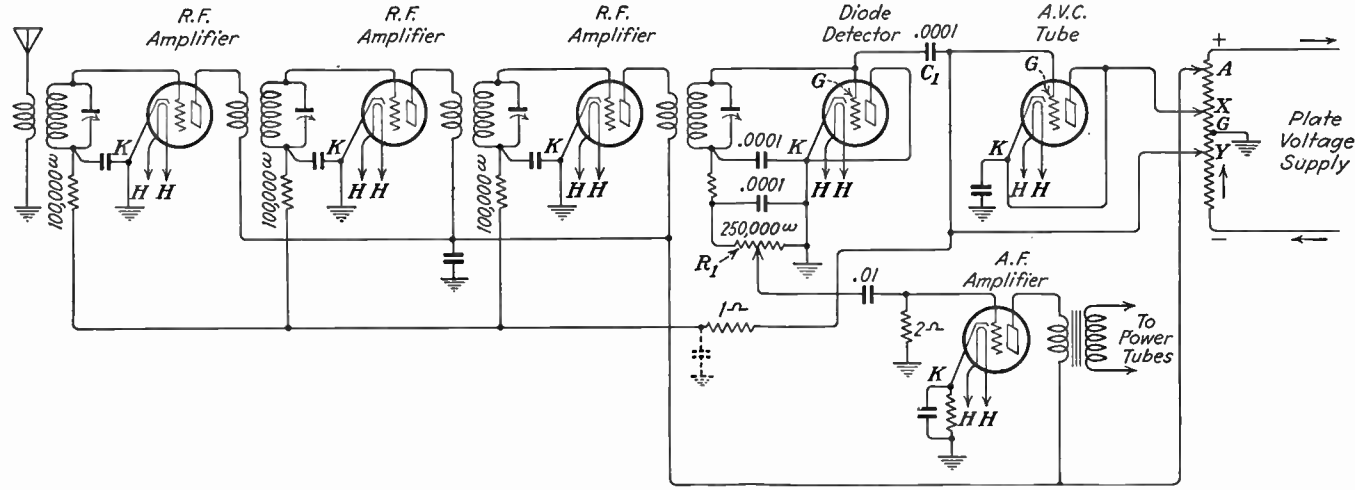


Fig. 161.—Delayed-action AVC circuit.

The theoretical operation of the automatic control is as follows:

Before a signal is applied to the grid of the first radio-frequency tube, it is assumed that all of the normal plate and grid-bias voltages are applied. When this is the condition, the plate voltages for all the tubes are obtained between the points *A* and *G* on the voltage divider. The normal bias voltage is obtained between the points *G* and *Y* of the voltage divider owing to the fact that the point *G* is the ground and all of the radio-frequency tubes have their respective cathodes connected to this point. Thus, since all of the grid-return leads connect to a point which is more negative than that of the cathode, namely, the point *Y*, the potential difference existing between these two points is the normal grid bias applied to all of the radio-frequency grids.

When a signal is applied to the grid of the first radio-frequency tube and amplified through the successive radio-frequency stages, a fairly large signal voltage may be developed across the input of the diode-detector circuit at the points *G*, *K*. This results in a rectifying action of the incoming modulated frequency in accordance with the diode theory of detection previously explained. The rectified-audio-frequency groups are then applied to the grid of the audio-frequency amplifier tube through the volume-control potentiometer *R*₁ and amplified at an audio frequency.

However, since the radio-frequency signal voltages developed across *G*, *K* of the detector tube are also applied across *G*, *K* of the *AVC* tube through the coupling condenser *C*₁, any voltage developed across the *AVC* tube will result in an increase in the potential across the voltage divider at the points *X*, *Y*. Hence, if the normal voltage existing between these two points is supplemented by an increased voltage, a current will flow in the *AVC* tube circuit and the point *Y* will be made still more negative than before, and an increased negative bias will result on the grids of all the radio-frequency tubes.

It may thus be readily seen that excessive signal voltages will raise the grid bias on the radio-frequency tubes through the *AVC* action, and thereby there will be maintained a fairly even signal output, free from overloading and consequent distortion at high signal peaks.

b. The most effective system of AVC action, generally known as delayed AVC, is one in which the control does not become operative until the signal voltage exceeds that of the divider at the points X, Y. That is to say, no current will flow in the AVC tube circuit until a voltage greater than that existing at these two points is developed. In this system the sensitivity of the receiver is greatly improved over the ordinary AVC circuit particularly on feeble signal voltages. In addition, the signal level is maintained at a much higher ratio with respect to the noise level inherent in high-gain receivers.

Noise suppression is, however, an independent action and is generally obtained by the use of a special tube possessing characteristics adaptable to this function.

c. The AVC Time Constant.—The degree of time required for a condenser to discharge its energy through a resistance is known as its *time constant* and depends upon the product of the capacity times the resistance. Thus,

$$T = \frac{CR}{10^6} \quad (123)$$

NOTE: The product must be divided by 10⁶ when *C* is expressed in microfarads.

For example, if a condenser of 0.01 mf. is charged and then connected across a resistance of 100,000 ohms, the discharge rate through the resistance will be $0.01 \times 100,000/1,000,000$, or 0.001 sec.

This is an important factor in efficient *AVC* systems since the delayed action maintains maximum sensitivity, provides maximum power output, and minimizes distortion due to excessive signal swing. In other words, the delay action provides automatic control only if the signal value exceeds a certain value, thereby enabling weak signals to be developed across the detector input without the piling-up effect of grid bias, and the resultant decrease in circuit sensitivity. Hence, only the powerful signals will be controlled by the delayed action, thus maintaining an average signal level without overloading.

The value of the time-delay constant for ideal performance should be between 0.04 and 0.09 sec. and should not be allowed to exceed the latter

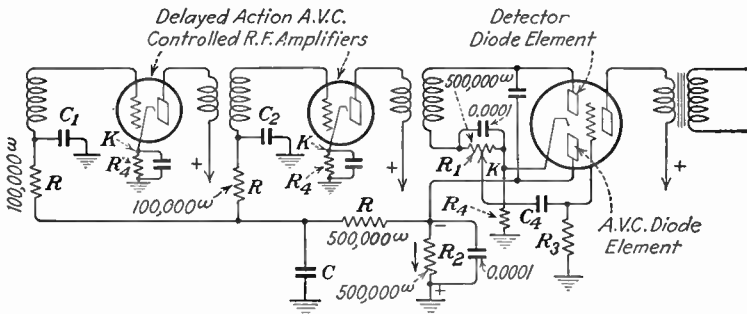


FIG. 162.—Diode detector and delayed AVC circuit.

value. If the constant is excessive, it will be impossible to provide automatic compensation for weak signals, particularly if they possess a fading characteristic, owing to the fact that the negative piling-up effect on the radio-frequency grids will be of excessive duration. Furthermore, much more difficulty will be experienced in tuning the circuit to resonance. An excessive time constant may readily be detected by adjusting the circuit to resonance with a powerful signal and with the volume-control set at the maximum gain position. If the circuit is quickly detuned by moving the tuning condensers to an off-resonance position the background noise in the receiver due to high-gain amplification should appear immediately. If a *QAVC* tube is used it should be opened when making this test. If the noise level builds up gradually, it is a definite indication that the time delay is excessive.

An explanation of time-delay action may be clearly understood by referring to Fig. 162. When an incoming signal is developed across the input of the diode-detector tube the asymmetrical or rectified variations will pass in direct-current pulses through the two resistances R_1 and R_2 . The rectified variations through R_1 are used to develop the audio-frequency voltages to be fed to the amplifier circuit in general conformity

to the theory of detection and audio-frequency amplification previously explained. However, since two complete diodes are used, the one connected to the resistance R_2 will also pass rectified-audio-frequency variations. These latter variations will be used to produce the automatic control action desired. Thus it may readily be seen that, as an incoming radio-frequency signal is rectified, a constant direct-current voltage will be developed across the resistance R_2 of a value depending primarily upon the amplitude of the signal voltage and the value of the resistance R_2 . Hence, the direct-current voltage across R_2 will charge the condensers C_1 , C_2 , and C to the same voltage as that being developed across R_2 . Consequently, if the incoming signal should possess a fading characteristic, the voltage developed across R_2 , while the signal developed is strong, will charge the condensers C_1 and C_2 and establish an increased bias on the two radio-frequency amplifier grids. However, if the signal intensity should suddenly drop due to a fading condition and the condensers C_1 and C_2 should have not discharged sufficiently through RR , R , and R_2 to release the throttling bias, the fading signal will not be compensated for and the automatic gain-control action will be lost. In other words, it is imperative for the condensers C_1 and C_2 to discharge fairly rapidly when the signal volume diminishes, to insure maximum sensitivity.

Although the main purpose of the resistances R_1 , R_2 , and R_3 and the condensers C , C_1 , and C_2 is for the purpose of grid filtration, that is, to prevent feed-back reactions between the various stages due to the common coupling connections, it is also important that the values of these condensers and resistances shall not impair the desired *AVC* action, particularly with fading signals. Hence, if the resistance values are made sufficiently large to prevent circuit reactions, yet not large enough to affect the circuit sensitivity owing to too much time lag, the operation of the circuit may be considered ideal for both sensitivity and automatic control. It is for this reason that the values of R and C must be properly proportioned in the grid-return circuit so that the time lag will not exceed the maximum 0.09-sec. value.

For example, let us assume that C , C_1 , and C_2 have a capacity of 0.015 mf. each and that the three filtering resistances, R , and the developing resistance, R_2 , have respective values of 100,000, 100,000, 500,000, and 500,000 ohms. This will introduce a total resistance of 1,200,000 ohms across the three condensers. The average time constant for this combination will be

$$T = \frac{CR}{10^6} = \frac{0.045 \times 1,200,000}{1,000,000} = 0.054 \text{ sec.}$$

Since this value is within the required range, satisfactory delay action should be obtained. It is important of course that the value of C and R for each grid filter be so apportioned that the radio-frequency voltage variations developed across R_1 will be dropped down to a negligible value across C_1 and C_2 .

148. Noise-suppression Circuit.—Noise-suppression control between assigned frequencies as effected by a tube and circuit arrangement designed for this purpose is commonly known as a *QAVC squelch*, or *Q-tube circuit*.

Figure 163 illustrates a practical arrangement of a noise-suppression circuit using a diode detector and a *QAVC* tube. In this particular system the first audio-frequency tube is automatically controlled, instead of controlling the radio-frequency amplifier as is done in the *AVC* circuit.

The use of this arrangement serves to reduce the inherent noises in modern high-gain receivers between the resonant adjustments of various stations. This is commonly known as interstation noise suppression.

The operation of this circuit, shown in Fig. 163, is as follows.

During the period in which no signals are being received and when no radio-frequency voltages are built up across the tuned circuit LC , the normal plate, screen, and bias voltages are applied to the two triple-grid tubes 78 and 77.

The plate of the 78 and the screen grid of the 77 tubes are tied together and receive their respective voltages from the voltage-divider system through the resistor R_1 . The normal grid bias for the 77 first-audio tube is supplied by the section of the voltage-divider system between the points X and Y .

When no signal voltages are applied across the tuned circuit LC , no voltage will be present on the control grid of the 78 $QAVC$ tube. This results in a large increase of the plate current for this tube owing to the large reduction in space charge because of the positively charged screen. At the same time this large plate current must be

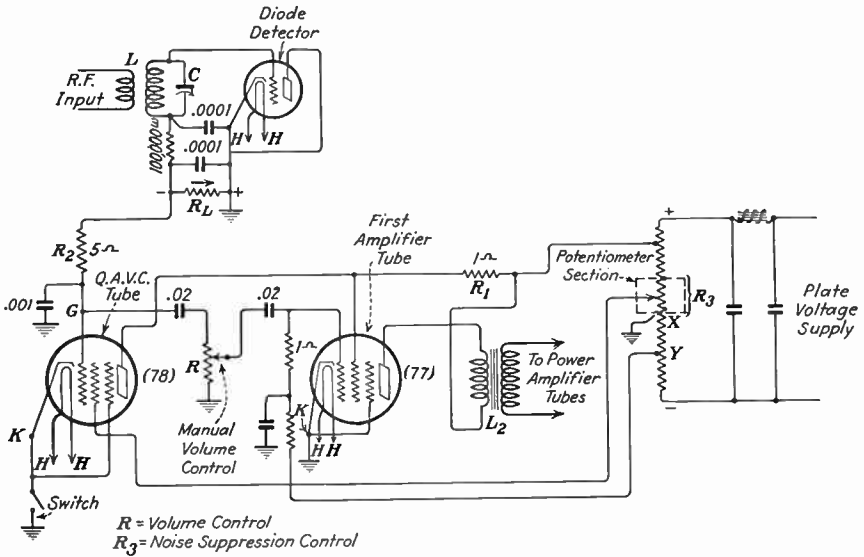


FIG. 163.—Inter-carrier noise-suppression circuit.

flowing through the resistance R_1 , resulting in a voltage drop across it. Now, since the screen of the audio-frequency amplifier tube 77 receives its voltage from this resistance, the IR drop across it will reduce this screen voltage. The design of this particular tube is such that a slight decrease in the screen voltage will greatly increase the space charge and reduce the plate-current flow. Consequently, during the period in which no radio-frequency signal voltages are being received, the first audio-frequency tube is operating with such little plate current (near cut-off) that no amplification can occur in the output or plate circuit. Thus no disturbing voltages developed in the radio-frequency stages can be transferred into the power-amplifier tubes and the loud-speaker system between the resonant adjustments of the tuning circuits.

However, as soon as the LC circuit and the preceding radio-frequency stages are tuned to resonance with a desired signal frequency, a voltage will be developed across the LC circuit of the diode detector, and pulsating (rectified) audio-frequency currents will be applied as low-frequency voltage variations, across the GK terminals of the 78 $QAVC$ tube. These low-frequency audio variations, owing to the diode rectifying action, are applied as groups of negative pulses to the grid of the $QAVC$ tube through the coupling and leak resistance R_2 . These pulses are of a negative character

by virtue of the signal current through R_L , as shown. In other words, the grid of the $QAVC$ tube will be negative with respect to the cathode K , which is at a positive potential with respect to the grid. The negative potential, applied as it is to the grid of the $QAVC$ tube, will reduce the plate-current flow in this tube, thereby reducing the voltage drop across the resistance R_1 . Hence the voltage increase on the screen of the first audio-frequency tube will restore the plate-current flow in its plate circuit and through L_2 . Thus the audio-frequency variations in the plate circuit in this tube will be transferred to the power amplifier and the loud-speaker system.

Hence, it may readily be seen that the 77 tube serves as a blocking device and will function only when radio-frequency voltages are built up across the LC circuit of the detector tube. Thus, since these voltages can only be established by a careful adjustment to resonance, all other disturbances between resonant adjustments will be rejected.

The $QAVC$ tube may be adjusted by the suppression-control potentiometer R_3 , so that the noise-suppression system will function for any signal level. In other words, if the screen voltage on the $QAVC$ tube is lowered by moving the sliding arm of the potentiometer down toward ground, the plate-current flow in the -77 tube will be increased, and, conversely, if the arm is moved upward, the screen voltage will increase and the cut-off point will be raised.

II. THEORY AND APPLICATION OF TUNED CIRCUITS

149. General Circuit Considerations.—The receiving circuit is fundamentally a generating device for the purpose of developing an e.m.f. of either radio or audio frequency to the grid of a vacuum tube. In addition, the circuit in itself may serve to increase the voltages to the grid of the tube through the medium of step-up transformers and properly designed L to C ratios. The circuit also may serve properly to segregate frequencies from one another (tune) by the variation of the circuit constants L , C and the mutual coupling between the circuits.

150. Theory of coupled circuits and tuning are the fundamentals of both the receiving and the transmitting circuits. It is essential that a knowledge of the alternating-current engineering principles covered in Chap. II be clearly understood, since these principles are directly applicable to all forms of receiving and transmitting circuits in which the problems of inductance, capacity, and resistance are involved.

The most common form of receiving circuit and its electrical equivalent is that illustrated in Fig. 164(a) and (b). Here the antenna and ground are shown connected to the primary winding of a radio-frequency transformer L_1 inductively coupled to a secondary winding L_2 which is tuned by a condenser C_1 .

The electrical equivalent [Fig. 164(b)] shows the component parts of the antenna-and-ground system which include the antenna inductance L_a , the antenna-to-ground capacity relationship C_a , and the entire antenna-circuit resistance R_a .

The antenna circuit, therefore, forms an oscillatory system which has resonant characteristics at a certain frequency in accordance with the ratio $f = \frac{1}{2\pi\sqrt{LC}}$. If the value of L, C is such that the resonant frequency is, for example, at 500 kilocycles, and the resistance R_a is negligible, the circuit would be effectively responsive only to the one frequency to which it is tuned, and consequently it would induce a *maximum* e.m.f. across the secondary circuit L_2C_1 at that one frequency only, provided, of course, that the secondary circuit is also tuned to the same frequency.

In other words, the circuit L_1 would have a *maximum* current flowing in it, only if the incoming-signal frequency is the same as that of the resonant period of the primary circuit. This would be a highly desirable condition if only one frequency were to be received, but, since in most

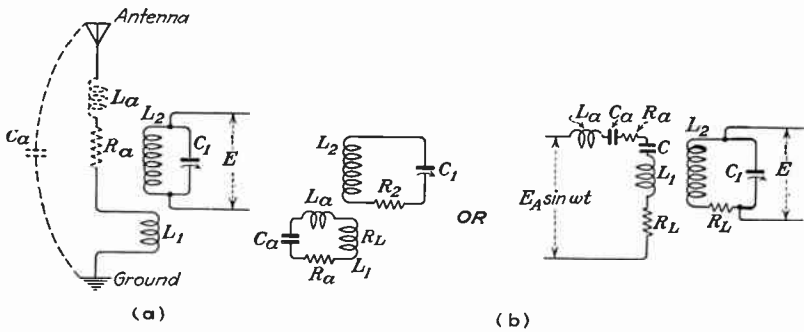


FIG. 164.—Simple receiving circuit and its electrical equivalent.

radio receivers the circuit must respond to a band of frequencies, the antenna circuit must be broadly responsive so that a number of radio frequencies are present in the form of a magnetic field around the primary winding L_1 . However, owing to the inherent direct-current and high-frequency resistance R_a in the antenna system, the sharply defined resonant characteristic is lost and the antenna circuit will respond to a broader band of frequencies such as is illustrated in Fig. 165. It will be noted that the resistance R_a is not included in the frequency equation, since resistance does not affect the frequency but merely tends to broaden the response and reduce the amplitude.

The broadening effect due to resistance in a resonant circuit is known as the *circuit damping* and is an appropriate expression for the energy dissipation in any circuit; that is to say, the more resistance introduced, the greater is the circuit loss, or damping. This is particularly true in the reception of modulated signals where the amplitude of the radio-frequency wave is varied at an audible rate, such as in broadcast and certain types of marine transmitters.

Circuit damping has both desirable and undesirable effects. In a broadcast transmitter, the frequency due to the modulation process,

produces a band of frequencies on either side of the tuned frequency, called side bands (Sec. 113). These bands extend to approximately 7,500 cycles on either side of the tuned frequency, or a band width of 15,000 cycles. Now since this band width includes the musical-frequency spectrum, it is essential that the circuit to receive these frequencies respond to at least a 15,000-cycle band if fidelity in musical reproduction is to be obtained. Consequently, a sharply tuned circuit would produce a selective response, as indicated in Fig. 165, which would result in the entire loss of the band frequencies or side bands. Serious distortion would obviously result.

However, when using unmodulated or continuous wave transmitters, the introduction of circuit resistance would merely reduce the amplitude of the signal voltage and reduce the overall sensitivity of the receiver.

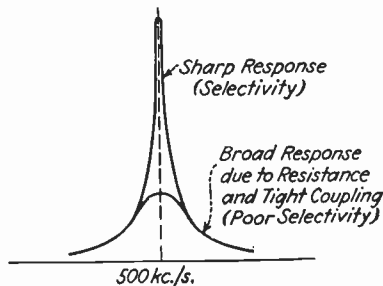


Fig. 165.—Sharp and broad tuning characteristics.

Hence, a sharply responsive or selectively tuned circuit is highly desirable for signals in which no side bands are involved and when a maximum amplitude is desired.

From this it may be seen that an antenna system which is to be sharply tuned must have a minimum of resistance, but, since this is practically impossible owing to the inherent coil and ground resistance, the alternative is to tune the antenna circuit for each frequency to be received. This can be accomplished by inserting a variable inductive or a capacitive reactance into the antenna circuit. If it is desired to tune to frequencies lower than the antenna natural-resonance period, an inductance coil is inserted, and, if it is desired to tune to frequencies higher than the antenna natural-resonance period, a variable condenser is inserted.

151. Simple Analysis of the Tuned Circuit.—It can readily be seen from Fig. 164(a) that the antenna circuit is equivalent to a series-resonant circuit, while the secondary circuit is of the parallel-resonant type. Consequently the impedance Z offered to the currents in both types of circuits will be quite different. This is an important factor since the voltages developed across any series or parallel combination will be dependent upon the impedance of the respective circuits.

In the primary portion of a receiving circuit, therefore, it is essential that a maximum current shall flow through the primary coil L_1 in order to develop a maximum field for the transfer of energy into the secondary circuit L_2C_1 . This condition is realized when the capacitive reactance X_c is equal to the inductive reactance X_L at some particular frequency.

The impedance of the circuit under these adjustments will be $Z = R$, since, series resonance is expressed:

$$|Z_o| = R^2 + \sqrt{(X_L - X_C)^2} \quad (124)$$

or,

$$Z = \frac{j\omega L + R + 1}{j\omega C}$$

Hence, since at resonance the impedance $Z = R$, the current flow in the circuit due to $E_A \sin \omega t$ will be entirely dependent upon the circuit resistance. If, therefore, this resistance is maintained as low as possible, a maximum current in the primary L_1 may be developed. The resistance R is primarily the high-frequency resistance in the coil, the antenna and earth losses, and the direct-current resistance due to the wire, which is dependent upon the number of turns and the size of the wire used.

The transfer of energy into the circuit $L_2 C_1$ may, under these conditions, be considered maximum and, if the circuit L_2, C_1 is tuned to resonance, a maximum voltage E will develop across the parallel mesh. Since it has been shown that the vacuum tube is primarily a voltage-amplifying device, this voltage must be made as large as possible so that the gain $\mu \times E_o$ through the tube will be maximum.

Hence the impedance of the parallel-resonant circuit should be infinite if a maximum e.m.f. is to be developed across it. Therefore, for parallel resonance

$$Z_o = \frac{X_c \sqrt{R_L^2 + X_L^2}}{\sqrt{R_L^2 + (X_L - X_C)^2}} \quad (125)$$

If the resistance of the coil is small compared to the reactance, as is usually true, the equation may be simplified to

$$Z = \frac{\omega^2 L^2}{R} \quad (\text{or formula 110}) \quad (126)$$

Thus a parallel-resonant circuit will approach an infinite impedance if the resistance R_2 is maintained as close to zero as possible. In other words, an increase of the coil resistance will decrease the impedance of the parallel combination.

Several other factors, however, will also affect the impedance relations between coupled circuits, particularly when signals of a modulated or damped character are applied to the primary circuit. The most important of these is the reactionary magnetic effects between the primary and secondary windings due to the mutual coupling between the circuits. For example, if the two coils L_1, L_2 are very tightly coupled, the energy developed in the secondary circuit will be partly dissipated in retransferring some of its energy back into the primary circuit owing to the mutual coupling. This is equivalent to increasing effectively the resistance R_2 of the secondary circuit $L_2 C_1$, so that the circuit impedance may be considerably reduced. In addition, the reflection of energy back into the primary circuit (called *reflected resistance*) will tend to raise the resistance of the primary circuit, so that the current flow in the series-resonant circuit will be reduced. The introduction of this resistance will increase the damping of the antenna circuit, which will tend to make it more broadly responsive, or less selective. Thus, since the resistances of both the primary and secondary circuits have been increased, the voltage E will be reduced. Therefore the broadening effect produced by the coupling reactions will make the circuit less selective, and interference from adjacent frequencies will result. One advantage, however, which will be derived from this condition is that the reception of modulated waves from broadcast transmitters will be more readily accommodated, or, in other words, the side-band frequencies will not be cut off and better fidelity will result.

Hence, the greater the mutual coupling (M) between two coils L_1 and L_2 , the greater will be the reactionary effects and the broader the tuning response.

$$M = \sqrt{L_1 L_2} \quad (42)$$

If the coupling between two coils is such that all of the lines of force produced by the primary and the secondary coils interlink, the percentage of coupling will be 100 per cent. But since this is practically impossible because of some flux leakage, the total coupling must be at some point below 100 per cent. This percentage may be calculated by taking the ratio of the mutual inductance (M) to the maximum theoretical value of L . This is called the *degree* or *coefficient of coupling* (k) and is expressed by the ratio

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (51)$$

From this, it may be readily seen that the mutual inductance between two coils will be maximum when the coils are tightly coupled and minimum when they are loosely coupled.

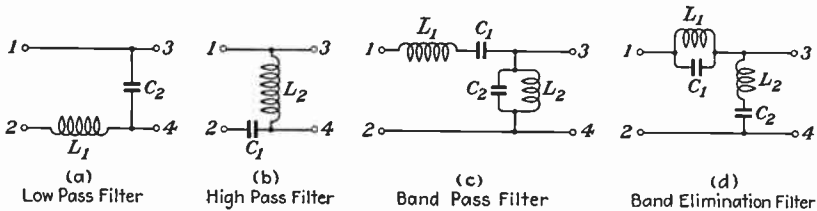


FIG. 166.—Filter circuits.

The modulated radio-frequency voltages E that are developed across any parallel mesh may then be applied to a suitable amplifying or rectifying device for either increasing the amplitude of these voltages or changing them into low-frequency asymmetrical pulsations (detection).

152. Filter Circuits.—In contrast to resonant circuits which are designed for operation at a definite resonant frequency, there is the filter circuit which provides attenuation or transmission over a definite slice of the frequency spectrum. Filters are used in radio-, audio-, and commercial-frequency circuits. When used in radio or audio circuits these circuits are referred to simply as *filters*, but, when used on commercial frequencies on power-supply apparatus, they are referred to as *ripple filters*. Ripple filters will be treated in detail in Chap. XIII on Rectifier Units.

For design purposes filters are divided into four classes, namely; low pass, high pass, band pass, and band elimination. These various types are shown in Fig. 166. The circuits shown are known as the *constant K type* because the effective impedance (or the geometric mean) of both the series and shunt sections remains constant over all the frequencies at which the filter operates, and this resistance is equal to the terminating resistance. The various classes of filters will now be discussed separately.

a. Low-pass Filters.—This form of filter is a circuit which is designed to attenuate all frequencies above a definite cut-off point (f_c) and to pass all frequencies below this point.

The low-pass filter may take the form of a simple- or complex-series or parallel-resonant circuit, or both. A simple L -type low-pass filter is shown as circuit (a). This circuit is derived as follows:

$$L_1 = \frac{R}{\pi f_c} \quad (127)$$

$$C_2 = \frac{1}{\pi f_c R} \quad (128)$$

where R is the resistance of the terminating impedance connected across 3 and 4. L and C are in henrys and farads, respectively. The cut-off frequency in cycles, f_c , is found by the equation

$$f_c = \frac{1}{\pi\sqrt{LC}} \quad (129)$$

b. High-pass Filters.—A circuit which is designed to pass all frequencies above a definite cut-off point and attenuate to a minimum value all frequencies below this point is called a *high-pass filter*. This circuit (b) may be derived as follows, using the same symbols as before:

$$L_2 = \frac{R}{4\pi f_c} \quad (130)$$

$$C_1 = \frac{1}{4\pi f_c R} \quad (131)$$

$$f_c = \frac{1}{4\pi\sqrt{LC}} \quad (132)$$

c. Band-pass Filters.—A circuit designed to pass a band of frequencies extending over a definite portion of the frequency spectrum and to attenuate all other frequencies is called a *band-pass filter*. A schematic diagram of such a circuit is shown at (c) and is derived as below. Let f_1 = lower cut-off frequency and f_2 = upper cut-off frequency. Then

$$L_1 = \frac{R}{\pi(f_2 - f_1)} \quad (133)$$

$$L_2 = \frac{(f_2 - f_1)R}{4\pi f_2 f_1} \quad (134)$$

$$C_1 = \frac{(f_2 - f_1)}{4\pi f_2 f_1 R} \quad (135)$$

$$C_2 = \frac{1}{\pi(f_2 - f_1)R} \quad (136)$$

$$f_1 = \frac{\sqrt{(L_2 + L_1)C_1} - \sqrt{L_2C_1}}{L_1C_1} \quad (137)$$

$$f_2 = \frac{\sqrt{(L_2 + L_1)C_1} + \sqrt{L_2C_1}}{L_1C_1} \quad (138)$$

d. Band-elimination Filters.—This type of circuit which is shown as (d) operates conversely to the band-pass filter in that it attenuates rather than passes a slice of the frequency spectrum. This circuit may be derived as follows, using the same symbols as before:

$$L_1 = \frac{(f_2 - f_1)R}{\pi f_1 f_2} \quad (139)$$

$$L_2 = \frac{R}{4\pi(f_1 - f_2)} \quad (140)$$

$$C_1 = \frac{1}{4\pi(f_2 - f_1)R} \quad (141)$$

$$C_2 = \frac{f_2 - f_1}{\pi R f_1 f_2} \quad (142)$$

153. Isolation and Band-pass Circuits.—Although the important consideration in all tuned-circuit designs is to develop a maximum voltage to the vacuum-tube amplifier or detector circuits, it has been shown that selectivity and broadness are also desirable. In the reception of broadcast signals both selectivity and broad signal response are required in order that interference may be eliminated and fidelity obtained. To accomplish this a number of tuned circuits must be employed to secure a result as shown in Fig. 167.

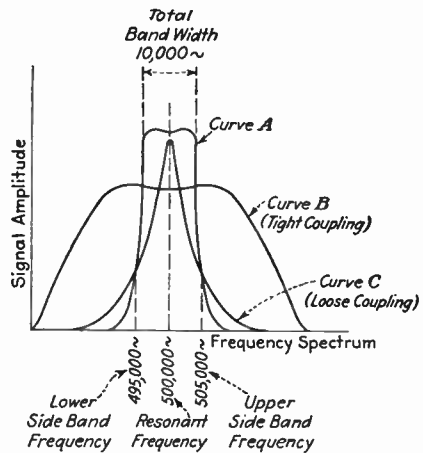


FIG. 167.—Curves illustrating band-pass effects.

Curve A shows the desired condition as obtained with a band-pass filter, while curve B shows the highly damped or broad characteristic generally found in tightly coupled circuits. Curve C shows the highly selective sharp response obtained with loosely coupled circuits.

The band-pass characteristic may be obtained by having two sharply tuned resonant circuits magnetically or electrostatically related so that their respective peaks combine into one to form the condition shown in (b), Fig. 168.

In (a), of the same figure, the two resonant peaks are separated to show the effect of two tuned circuits not coupled to each other. If now these circuits are properly combined through coupling, the two curves may be made to produce the various responses illustrated in (b). This depends primarily upon the ratio of the inductive reactance X_L to the resistance of the circuits R and is expressed X_L/R which is generally referred to as the Q of the coil or the circuit. If the ratio is extremely high or approaching infinity, the two peaks will resemble curve 1. If the ratio is too low, the curves will flatten out and produce a response as

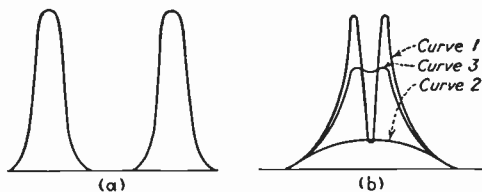


FIG. 168.—Curves illustrating individual and combined coupling reactions.

in curve 2. Hence by the proper manipulation of the circuit Q , an effect in between curves 1 and 2 may be obtained as illustrated by curve 3.

Three typical arrangements for obtaining band-pass effects are illustrated in Fig. 169(a), (b), and (c).

The two resonant frequencies in (a) are obtained by adjusting the two circuits L_1C_1 and L_2C_2 to the desired frequency so that two separate peaks are developed across each branch. If the coupling M is assumed to be very loose, the two response curves will be as indicated in curve 1, Fig. 168. However, as the coupling M is increased the coupling Q will

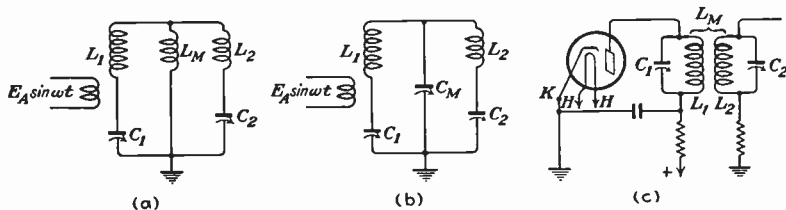


FIG. 169.—Band-pass filter circuits.

decrease and approach the condition of curve 3, Fig. 168. Hence, as the coupling impedance due to the increase in Q goes up, the curves will eventually combine to form the band-pass condition illustrated in curve A, Fig. 167.

The coil L_M in Fig. 169(a) is not a physical body but serves merely to represent the magnetic coupling between L_1 and L_2 . Similarly, in Fig. 169(b), the two coils are coupled through the medium of an electrostatic coupling in which C_M represents the mutual effect between the coils, due to a capacity. If the capacity of the condenser C_M is increased, thereby decreasing its reactance, the two tuned circuits L_1C_1 and L_2C_2 are loosely

coupled to each other and produce a response equivalent to the loose coupling condition due to L_M in the magnetically coupled circuit. Obviously, if this capacity is decreased, the increase in its reactance will serve to combine the two peaks as in the case of a tightly coupled magnetic circuit in Fig. 168(b).

Perhaps the simplest and most practical application of the band-pass filter is that of the two tuned parallel-resonant circuits illustrated in Fig. 169(c). When the two circuits L_1C_1 and L_2C_2 are tuned to their respective resonant frequencies with a minimum degree of coupling L_M , the two response curves will represent the condition of Fig. 168(a). Then, as the coils are brought closer together, thereby increasing the coupling, the two peaked curves are merged and broadened to give the desired band-pass condition of curve A, Fig. 167.

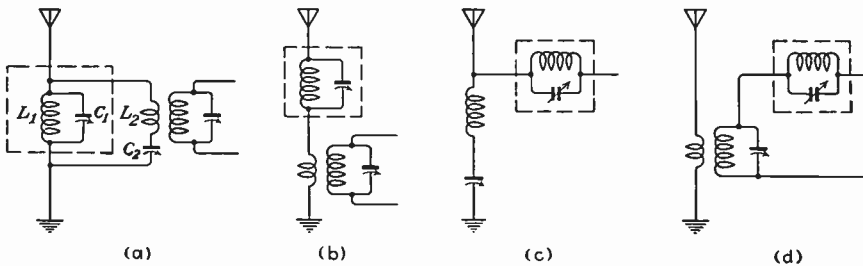


FIG. 170.—Rejector and acceptor circuits (wave traps).

Although the band-pass filter produces the desired result in accommodating side-band frequencies, it presents a disadvantage in that the currents circulating in the various-tuned circuits produce a dissipation of the signal energy which ultimately reduces the signal voltages applied to the radio-frequency amplifier tubes. This disadvantage, however, may be overcome by introducing an additional stage of radio-frequency amplification using a screen-grid tube of the high- μ variety. Hence, from the expression, $\text{gain} = \mu \times E_o$, it will be seen that a decrease in signal voltage E_o due to the band-pass circuit can be compensated for by an increase in the amplification factor (μ) of the tube.

154. Rejector or Trap Circuits.—It is frequently necessary in receiving or transmitting systems to introduce an additional tuned impedance which will offer to the desired frequency a high-reactance path, while to undesired frequencies it will provide a low-reactance path. In this manner it will be possible to have two tuned circuits functioning simultaneously so as to accept or reject certain frequencies, as desired.

Figure 170(a) illustrates a typical trap circuit which may be designed to eliminate undesired frequencies by adjusting the parallel-resonant circuit L_1C_1 to that of the desired frequency while the values of L_2 and C_2 are also adjusted to the desired frequency.

It has previously been shown that the impedance of a parallel-resonant circuit, when adjusted to resonance, will be

$$Z_{p.r.} = \frac{L}{CR} \quad \text{or} \quad \left(\frac{X_L^2}{R} \right) \quad (143)$$

while that for a series-resonant circuit at resonance will be

$$Z_{s.r.} = \sqrt{R^2 + (X_L - X_C)^2} = Z = R \quad (73)$$

If, therefore, the two circuits are critically adjusted so that the ratio L/CR of L_1C_1 offers a high impedance to the desired signal frequency, then the maximum signal currents will pass through L_2C_2 . The undesired frequency, on the other hand, will not pass through L_2C_2 but will pass more readily through L_1C_1 . It is, of course, possible that some of the

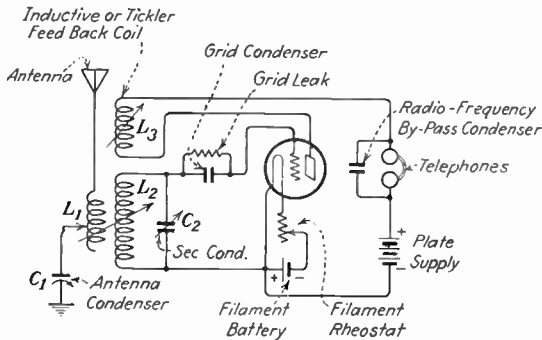


FIG. 171.—Regenerative receiver circuit.

interfering signal may pass through L_2C_2 , and it is, therefore, necessary to arrange the relating impedances so that $Z_{s.r.}$ is high compared with $Z_{p.r.}$ at the undesired frequency.

The efficiency of the rejector circuit will depend greatly upon the accuracy of its adjustment and its resistance value. It is imperative that the coil L_1 possess a high Q characteristic, otherwise the degree of shunting effect across L_2C_2 to the undesired frequency may be greatly reduced.

Various other forms of rejector circuits (enclosed by dotted lines) are illustrated in Fig. 170(b), (c), and (d).

155. Regenerative Receiver.—It will be remembered that the rectifying action of the tube, due to the grid condenser and leak, resulted in a low- and a high-frequency current flow in the output-plate circuit, in which the low-frequency pulse passed through the headphones (telephones) and the high-frequency pulse was by-passed through the fixed condenser, shunting the headphones.

In a regenerative system it is this radio-frequency current in the plate circuit of the detector which can be used for radio-frequency amplification

of the received signals by a feed back of energy to the grid of this tube through either an inductive or capacitive coupling.

Figure 171 illustrates a typical regenerative circuit in which the coil L_3 is used to feed back the signal radio-frequency variations in the plate circuit to the grid to produce an increased signal variation and, consequently, a louder response in the headphones.

The explanation of this action is as follows:

In the high-frequency grid circuit L_2C_2 , where the damped-signal variations are fed to the grid after they have been induced by conditions of resonance from L_1C_1 , there is always a certain amount of ohmic resistance present, which results in a loss of some of the signal energy, and, consequently, an increase in the signal damping. This loss, due to the circuit resistance, is called the *positive resistance* of the circuit.

When the signal variations are applied to the grid of the regenerative receiver, there will be a resultant high-frequency plate-current component which will flow through the coil L_3 and the radio-frequency by-pass condenser. Now, if the coil L_3 is inductively related to L_2 , an e.m.f. will be induced across L_2C_2 which will be in phase with the high-frequency e.m.f. set up in the same circuit by the incoming signal. If, then, the coil L_3 is wound so as to provide the proper feed-back coupling between the plate and the grid circuits, the positive ohmic resistance of the circuit will be partly neutralized, which will result in a decrease of damping and therefore a higher amplitude of the signal variations and, consequently, a louder signal response in the telephones. The feed-back energy from the plate to the grid circuit acts to neutralize the effect of the positive resistance and is called *negative resistance*.

It is important to note that the neutralizing effect increases as the feed-back coupling L_3 is made closer, and a point may be reached where the damping is so small, owing to the neutralization of the positive resistance by the negative resistance, that the circuit will start oscillating. An incoming modulated or damped signal will by virtue of its sharply tuned character have its side-band frequencies narrowed or cut off, if the circuit is oscillating, and the signal will be heavily distorted and possibly inaudible.

It can thus be seen that a regenerative system will give increased amplification of damped and modulated signals only up to a certain point, that is, a little below the oscillating point.

If, on the other hand, however, incoming signals of an undamped character (*CW*) are applied to the grid of the regenerative detector, no signals will be heard until the tickler coil is adjusted to the point of self-oscillation. When the regenerative detector is brought to the point of self-oscillation, the sustained voltage variations on the grid due to oscillation are combined with the incoming-signal voltages, which produces a resultant "beat" in the plate circuit which is the difference between the two sustained frequencies. Hence, if the incoming-signal frequency is 500 kilocycles, and the tickler coil is so adjusted that the circuit will self-oscillate at 501 kilocycles, the resultant beat frequency will be 1,000 cycles. Consequently, since this beat frequency is in the audio-frequency range, low-frequency variations will pass through the telephone circuit which will render them audible. By the critical adjustment of the tickler coil, the self-oscillating frequency may be varied to change the resultant beat frequency to any desired audio-frequency tone between 500 and 5,000 cycles.

This system whereby a single tube functions as an oscillator and a detector is the most commonly used for the reception of telegraphic continuous-wave signals and is generally known as the *autodyne* method of beat reception.

If a receiver with a non-regenerative detector is used no *CW* signals can be received unless a separate or external oscillator is provided to beat against the incoming signal. The oscillator is generally magnetically coupled to the detector circuit.

It is readily seen that a regenerative detector used as an autodyne, that is, in an oscillatory condition, will radiate energy through the antenna system if connected as in Fig. 171. Usually this radiation is of such a feeble character as to cause no serious interference. However, in order to prevent the autodyne from causing interference, it is the usual practice, in modern receivers utilizing this type of detector, to precede the detector with one or two stages of radio-frequency amplification. The radio-frequency tube (or tubes) then acts as a *buffer tube* and prevents energy generated in the autodyne tube from reaching the antenna where it would be radiated as an interference wave.

156. Superregenerative Receiver.—It has been shown in the regenerative circuit that the sensitivity of the receiver is dependent upon the introduction of sufficient feed back from the plate to the grid circuit to introduce an e.m.f. which will reduce the inherent circuit-resistance losses and thereby increase the amplitude of the voltage applied to the grid of the detector tube. These voltages are increased up to a critical point just before the tube goes into self-oscillation.

In the superregenerative receiver it is possible to increase these voltages to a larger amplitude before the tube goes into self-oscillation, thereby greatly increasing the amplification or sensitivity of the receiver.

These conditions are obtained by coupling a local oscillator to the normal regenerative circuit and adjusting the local oscillator to the point in which the damping of the regenerative circuit is periodically increased. This local generating frequency must be of a relatively low value, say 10 to 25,000 cycles so that the incoming modulated signal frequency when supplemented by the oscillator frequency is just below the point of self-oscillation.

Figure 172 illustrates a circuit arrangement of a highly sensitive superregenerative receiver for the reception of ultra-high frequencies.

The principal reason for the use of the superregenerative receiver circuit on ultra-high-frequency systems is because it tunes very broadly owing to the modulation of the detector by the local oscillator, resulting in a selectivity characteristic several hundred kilocycles wide. Such a receiver is a satisfactory counterpart of a communication system utilizing a self-controlled oscillator, which is consequently subject to considerable frequency creeping. If a more selective circuit such as a straight regenerative or superheterodyne were employed to receive emissions from a self-controlled oscillator, a considerable loss of side-band response would result causing a poor quality in the received radio-telephone signal, or, if telegraphy is employed, it would be necessary for the operator continu-

ally to adjust his receiver to follow the frequency variation of the transmitter. When stable transmitters are employed a more selective receiver will be desirable to do its part in cutting out interference.

157. Superheterodyne or Double Detection Receiver. *a. General Considerations.*—It has been shown that the gain in a radio-frequency amplifier depends upon several general factors, namely, the efficiency of the circuit design, the number of tubes used, the μ of the tubes, and the ratio of turns in the primary and secondary windings of the radio-frequency transformer.

The voltage developed across the primary winding of a radio-frequency transformer is found by the equation

$$E_p = \frac{\mu E_0 Z_p}{R_p + Z_p} \text{ volts} \tag{118}$$

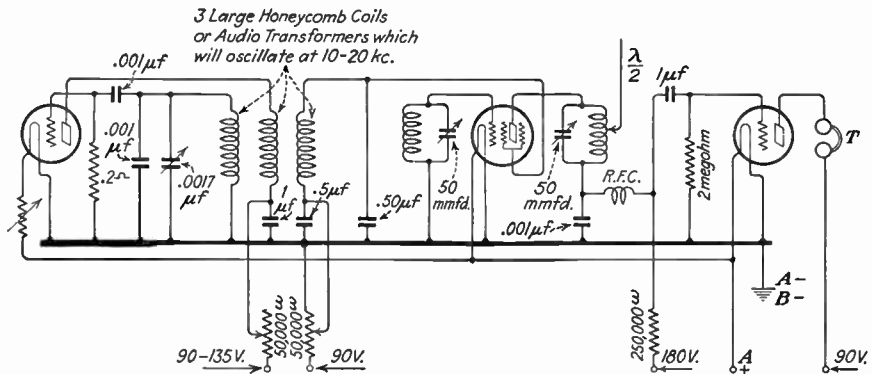


FIG. 172.—Superregenerative receiver circuit for ultra-high-frequency reception.

The voltage developed across the secondary of a radio-frequency transformer is found to be

$$E_s = \frac{\mu E_0 Z_p}{R_p + Z_p} \times \frac{L_s}{M} \text{ volts (originally formula 102)} \tag{144}$$

From this it may be seen that the gain of a single radio-frequency stage could be greatly increased by using a radio-frequency transformer with a very large turns ratio. This is, of course, quite impossible in a radio-frequency amplifier covering the broadcast band, since the large increase of L would reduce the frequency period of the circuit and make it possible to tune to low-frequency bands only.

If, however, a decreased frequency of say 175,000 cycles is passed through a radio-frequency transformer having a large turns ratio, a considerably greater voltage would be applied to the succeeding amplifier tube, resulting in a greater gain, E_2/E_1 . It is this principle which makes a single-stage, superheterodyne intermediate-frequency amplifier a relatively higher voltage-amplifying device, and thereby greatly increases its sensitivity.

Figure 173 illustrates a block arrangement of the component units of a superheterodyne receiver as used for the reception of modulated or damped signals. The use of this circuit for the reception of undamped continuous waves (*CW*) (unmodulated) will be explained later.

The high-frequency unit may consist of a number of radio-frequency amplifier stages tuned to the particular frequency desired. These frequencies are then applied to a detector or mixer tube in which the signals are mixed with the local-oscillator frequencies and thus changed from a high to a lower frequency signal, so that they may be passed through a radio-frequency transformer of lower frequency characteristics. This type of transformer has an increased step-up ratio and, consequently, will develop larger voltages to be passed on to the succeeding tube, the intermediate radio-frequency amplifier.

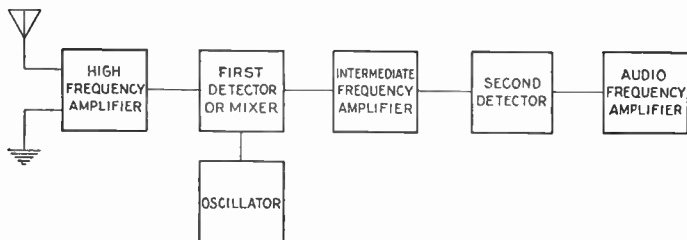


FIG. 173.—Block diagram of superheterodyne system.

The oscillator tube generates sustained oscillations which are also applied to the first detector tube. These oscillations are of such value that, when they are reacting with the incoming-signal voltage on the grid or plate of the first detector tube, they produce a wave which is the difference between the incoming signal and the oscillator frequency. If this difference, or beat frequency lies somewhere in the lower frequency range, say 175 kilocycles, and the output transformer of the first detector tube is tuned to this frequency, the resultant signal will be applied to the grid of the succeeding tube. This latter tube is known as the intermediate-frequency amplifier. A number of these stages may be used depending upon the sensitivity desired.

From here the signal is applied to the second detector tube where it is rectified and amplified in the usual manner.

Figure 174 illustrates a complete superheterodyne receiver using screen-grid tubes of a high-gain type and efficient circuit characteristics. These principles apply to all types of superheterodyne receivers as may readily be seen by referring to some of the commercial types used in aircraft and marine communication.

b. Theoretical Operation of the Receiver for Modulated Waves.—Assuming that all plate, screen grid, and grid-bias voltages are applied, and that the tube filaments are lighted, the operation of the complete receiver is as follows.

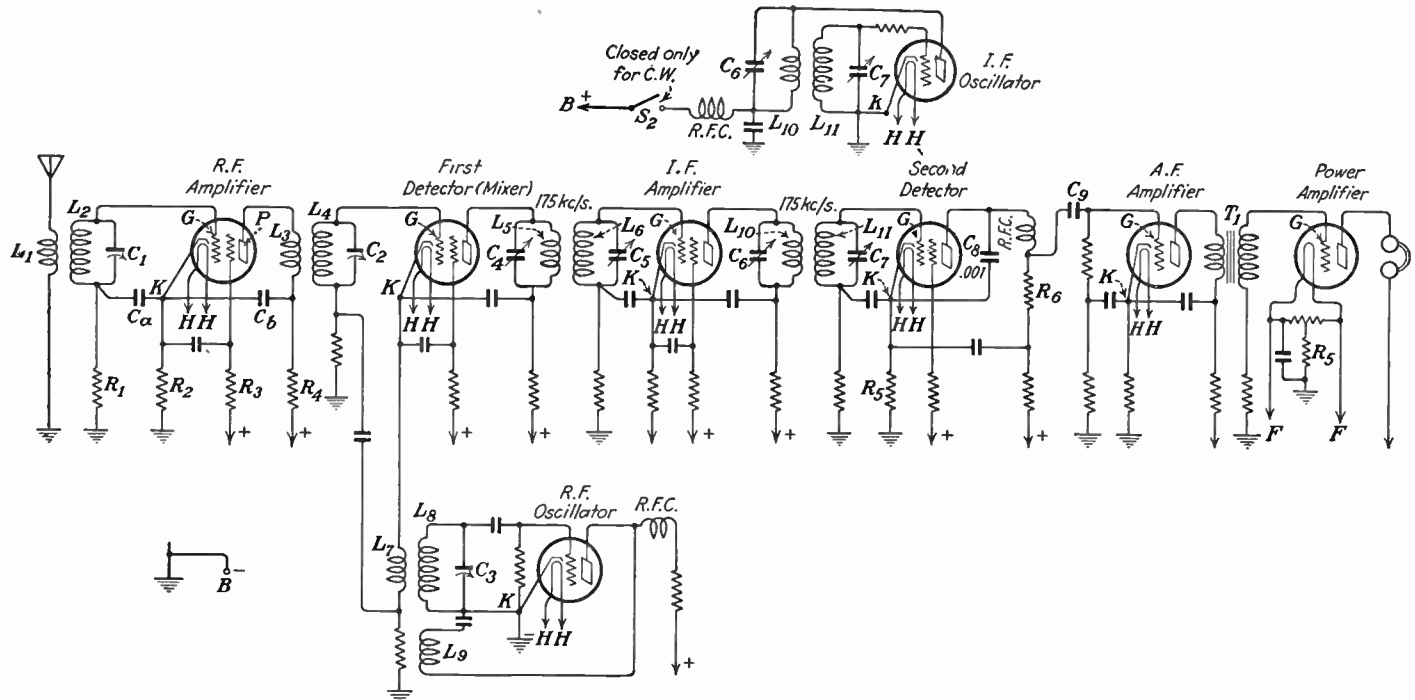


FIG. 174.—Superheterodyne receiver circuit.

When a signal voltage is developed across L_1 , an alternating current of a definite frequency will flow in the antenna circuit. This will produce a varying magnetic field around the winding and induce an e.m.f. across the secondary circuit L_2C_1 , if the circuit is adjusted to resonance. The voltage developed across the parallel mesh will then be applied between G and K of the radio-frequency amplifier through the by-pass condenser C_a . The signal voltage E_g developed across G and K will be applied across the load impedance L_3 of an approximate value, μE_g . This results in a radio-frequency current variation in L_3 and C_b . These variations are isolated from the supply circuit by the plate-filter resistance R_4 , thereby preventing magnetic interactions with other circuits common to the supply. This increases the efficiency and the stability of the amplifier. The grid filter and condenser, R_1 and C_a , respectively, serve a similar purpose in the input or grid circuit (see Plate and Grid Filtration, Sec. 145).

The magnetic variations now present across L_3 are induced into L_4 . The increased signal voltage now present across G and K of the mixer tube develops a radio-frequency variation in the load circuit C_4L_5 . However, since this circuit is resonant to a frequency of 175 kilocycles and the incoming frequency is in the broadcast spectrum of between 500 and 1,500 kilocycles, depending upon the adjustments of L_2C_1 , L_4C_2 , no e.m.f. will be developed across the secondary circuit, L_6C_3 .

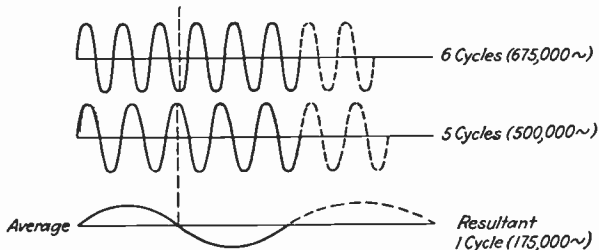


FIG. 175.—Circuit reactions in a beat receiver.

Let us assume the incoming radio signal to be at a frequency of 500 kilocycles. In order to transfer this signal through the 175-kilocycle transformer, the radio-frequency oscillator must be set into operation. The oscillator is of the inductive-feedback variety and operates upon the fundamental principles of the vacuum-tube oscillator explained in the preceding chapter.

The condenser C_3 is now adjusted so that the oscillator will generate a radio frequency which, when combined with the incoming-signal frequency of 500 kilocycles, will produce a beat frequency of 175 kilocycles.

With the radio-frequency oscillator assumed to be generating a frequency of either 675 or 325 kilocycles, a voltage in proportion to these frequencies will be induced across the coil L_7 in series with the cathode of the first detector tube. Now, since there are already signal voltages developed across L_7 by the incoming signal, the combining of these two voltages in this circuit will generate an e.m.f. and a resultant beat frequency, which is the difference between the two, 675 - 500 or 175 kilocycles. This condition is illustrated in Fig. 175. Frequencies of 6 and 5 cycles are shown in the illustration for the sake of simplicity.

This frequency is then induced into the 175-kilocycle resonant circuit, and an e.m.f. is developed across G and K of the intermediate-frequency amplifier tube. This frequency is amplified again through the second 175-kilocycle radio-frequency transformer L_{10} , L_{11} , where it is applied across G and K of the second detector tube. This tube is biased near the cut-off point by the biasing resistor R_5 , to produce the necessary rectifying action for detection of the 175-kilocycle modulated signal to effect audibility. The radio-frequency asymmetrical variations now present in the

plate circuit develop group voltages across the by-pass condenser C_8 , which are discharged as low frequency or audio pulses through the plate-coupling resistance R_6 . This resistance is used as the coupling medium instead of an audio-frequency transformer because of the high plate-resistance load required to match the high screen-grid tube impedance R_p .

The audio-frequency voltages now developed across R_6 and the plate coupling condenser C_9 are applied to G and K of the audio-frequency amplifier. The audio-frequency variations now present in the plate circuit of this tube are applied to the grid of the power tube through an audio-frequency transformer T_1 .

The bias voltages for all tubes are obtained from the cathode resistances connected in series with each cathode. In the power tube a resistance R_8 provides the grid bias since no cathode-heater-type tube is used.

In addition to the high degree of amplification which may be obtained from the superheterodyne receiver, it possesses exceptional selective qualities owing to the beating effect of the oscillator with the incoming radio-frequency signal.

One important difficulty encountered in the superheterodyne receiver is the interference caused in the radio-frequency stages owing to the reaction of the oscillator with certain other incoming-signal frequencies.

For example, let us assume that a signal frequency of 2,000 kc./s. is to be received. The oscillator may be adjusted to either 2,175 or 1,825 kc./s., since the resultant beat in either case will be 175 kc./s. Now, if a signal frequency of 2,350 kc./s. is also being received, the difference between 2,350 and 2,175 kc./s. is also 175 kc./s. This is known as the *image frequency*. This interference may be eliminated by loosening the coupling in the radio-frequency transformers L_1 , L_2 and L_3 , L_4 (Fig. 174).

Another method of suppressing the image frequencies would be to connect a parallel-resonant circuit in series with the grid of the first detector tube, and adjust the circuit so that it introduces infinite impedance to the image frequency. This may be accomplished by adjusting this circuit so that it will differ from the desired signal by twice the intermediate frequency, under which conditions no image frequencies will be introduced beyond the mixer circuit.

c. The Superheterodyne for Continuous-wave (CW) Reception.—If it is desired to receive continuous waves with the superheterodyne receiver, the second detector must employ the autodyne circuit, or an external oscillator must be coupled to the grid of this detector tube. The latter method is usually employed, and the external oscillator so used is referred to variously as the second oscillator or as the intermediate-frequency (*IF*) oscillator. Regardless of the system used, the circuits between the oscillator and the intermediate-frequency amplifier should be carefully shielded to avoid circuit instability.

Theoretical Operation: Let it be assumed that the *IF* circuit of the receiver is tuned to 175 kc./s. An incoming *CW* signal applied to the grid of the first detector or mixer tube, as shown in Fig. 174, will pass through the intermediate stage owing to the heterodyning effect of the first oscillator tube in accordance with the superheterodyne principle explained in *b*.

This signal eventually reaches the second detector tube, but it is still a continuous wave of an inaudible frequency (175 kc./s.). In order to make the signal audible, an *IF* oscillator, previously referred to, is coupled to the second detector and adjusted to beat against the 175-kc./s. *IF* signal. This produces an audible beat note equal in frequency to the difference between the *IF* signal and the *IF* oscillator frequency.

Thus to get a 1,000-cycle note, the *IF* oscillator is adjusted 1,000 cycles either side of the 175-kc./s. *IF* signal; that is, to 174 or 176 kc./s.

158. The Single-signal Receiver.—Although the superheterodyne receiver provides excellent selectivity on the broadcast band, it has been found greatly lacking in this respect on the very high frequencies in continuous-wave reception, owing to the crowding of stations into very narrow bands. This is particularly true in the short-wave-amateur bands, and in the various marine bands in which *CW* code transmission is used.

The inherent qualities of a quartz plate, which cause it to provide extremely selective tuning because of its sharp resonant characteristics

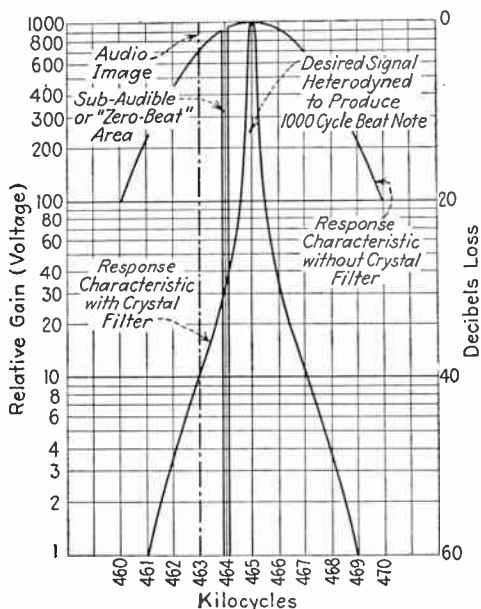


FIG. 176.—Response curves with and without a quartz-plate filter. (Courtesy of Hammarlund Manufacturing Company.)

at certain frequencies, when operated at a fixed temperature, make it highly desirable for use as a filtering system in a receiving circuit to provide extreme selectivity, particularly in crowded high-frequency spectrums where rigid discrimination between stations is essential.

The characteristics of a receiver using a *crystal filter* are such that when the circuit is adjusted to approximately 1,000 cycles off the resonant point, the voltage gain of an interfering signal may be reduced as much as 97 per cent as shown in Fig. 176. Comparing this with the results obtained with a superheterodyne receiver without a crystal filter shows that in the latter the interfering signal, 1,000 cycles off resonance, is reduced only 5 per cent. This indicates that the crystal filter is highly desirable for the reception of the unmodulated or *CW* signal, such as is

used in code transmission; but that it is obviously undesirable for broadcast frequencies where the frequency band must be 10 kilocycles wide to accommodate the side bands in the musical spectrum.

It has been previously shown that the greatest difficulty encountered in the superheterodyne receiver, however, is the interference due to the so called *image frequency*. This image frequency will be particularly objectionable on crowded *CW* bands, owing to the close proximity of stations. With the crystal filter, however, it is seen on the chart that the interfering image-frequency signal may be reduced to a negligible quantity (99 per cent) before it reaches the *IF* amplifier circuit.

Theoretical Operation: It has already been shown how the *IF* oscillator in the superheterodyne receiver serves to make *CW* reception audible. However, other factors enter into the problem when the receiver is used on crowded frequency channels, resulting sometimes in serious interference. Such a situation will now be explained.

Suppose station *A* transmitting on a frequency of 6,000 kc./s. is being received. This means that the radio-frequency oscillator of the receiver must be adjusted to

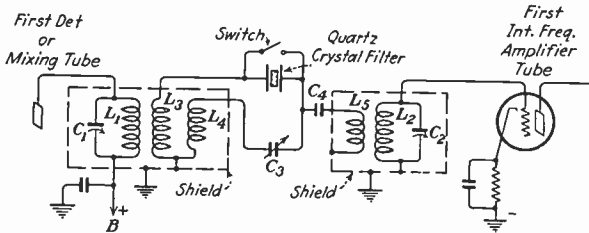


FIG. 177.—Quartz-plate filter circuit.

produce an intermediate beat frequency of 175 kc./s., assuming this frequency, as usual, is the *IF* of the receiver. Thus the radio-frequency oscillator must be adjusted to 5,825 or 6,175 kc./s.

Now suppose station *B*, capable of radiating an interfering signal, is transmitting on a frequency of 5,998 kc./s. which is 2 kc./s. lower than the desired signal. Thus when the emissions of *A* and *B* interact, a heterodyne frequency of 2 kc./s. is generated.

There are then present, in addition to the intermediate frequency of 175 kc./s. two other frequencies, (1) the 2 kc./s. and (2) the radio-frequency oscillator frequency which can be assumed to be 6,175 kc./s. These frequencies heterodyne, with the result that frequencies to the value of 6,173 and 6,177 kc./s. are generated, resulting in two new intermediate frequencies of 173 and 177 kc./s. These are very close to the resonant frequency of the *IF* amplifier, and consequently a considerable amount of energy gets through the *IF* stage. These interfering frequencies are capable of producing the same effect on the second detector as the desired signal; they are the *image*, so to speak, of the desired signal, and they are, therefore, called *audio-image frequencies*. The reason for the prefix *audio* is that they produce an audio response, by heterodyning or beating with *IF* oscillator, similar to that of the desired signal.

Obviously, if these interfering frequencies could be prevented from reaching the *IF* amplifier in the first place, they could not finally reach the second detector and result in audio-image frequencies. It is to this function that the crystal filter, as used with the superheterodyne single-signal receiver, lends itself. The method of introducing the filter into the superheterodyne circuit is shown in Fig. 177.

The connection diagram is conventional, except for the provision made for the introduction of the crystal filter. The primary L_1 and the secondary L_2 of the coupling transformer are placed in separate shielded compartments. The split windings L_3, L_4 are inductively coupled to L_1 , thereby providing for an input circuit to the crystal filter. The output of the crystal is inductively coupled to the secondary L_2 through C_4 and L_5 . At resonance, L_1 and L_3 , and L_5 and L_2 are matched impedances.

The coil L_4 and the condenser C_3 neutralize the capacity of the crystal holder plates. Coils L_3 and L_4 have induced in them e.m.fs. which are 180 degrees out of phase. Therefore, when C_3 equals the capacity of the crystal-holder plates, any currents which these plates might by-pass through their capacity are canceled out by currents in opposite phase fed through C_3 . This leaves only the signal e.m.fs. passed by the crystal free to reach the input circuit of the *IF* amplifier.

The setting of C_3 , therefore, governs the selectivity of the crystal, and the band passed through the crystal filter can be slightly widened, enough to make speech intelligible, but not enough for broadcasting which requires a much wider band. This condenser for this reason is sometimes known as the *interference control*.

This system is of great value in making the handling of traffic possible in what would otherwise be an overcrowded spectrum. It promises soon to be almost indispensable on receivers used in certain frequency bands.

CHAPTER VI

ANTENNAS AND WAVE PROPAGATION

In order to radiate electromagnetic and electrostatic energy into space, it is necessary to obtain a circuit which will not confine the useful fields to the immediate vicinity of the coil and condenser constituting the closed oscillatory circuit, where it would be absorbed. It is found, on the other hand, that an *open oscillatory circuit*, which has its inductance and capacity distributed over a large area will radiate a large part of the energy flowing in the system.

I. FUNDAMENTAL TYPES OF ANTENNAS

159. The Half-wave Doublet.—The first radiators used by Hertz were a pair of short brass rods. They were called *doublets*, but were not a half wave length long. Many of the doublets used in directive arrays

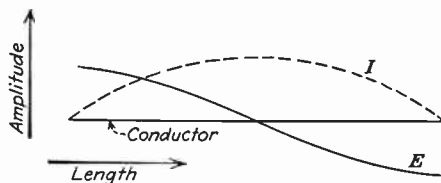


FIG. 178.—Current and voltage relations in a half-wave doublet.

(antenna systems) today are not a half wave long, so that the term *doublet* should not be used indiscriminately to mean a half-wave antenna.

The half-wave doublet, however, may be considered as the basis from which most other types of antennas are developed. It consists of a conductor approximately one-half wave length long, which is excited by some coupling system from an oscillator or a radio transmitter. The method of exciting the half-wave-length doublet does not change the voltage and current distribution curves, so the coupling circuits will be considered in a later paragraph. These voltage and current curves are shown in Fig. 178. In this diagram the voltage (curve *E*) is the voltage with respect to ground, or to the center of the radiator. The current shown (curve *I*) is that which would be found passing a given point, if a cross section of the conductor were taken, and could be measured by inserting a thermogalvanometer at successive distances from the center. This open oscillatory circuit functions in accordance with the theory of parallel resonance, the same as a coil-condenser combination. The wire along its length has inductance, and the capacity between one *pole* and the other constitutes

the capacity of the resonant circuit. The current in all parts of the conductor is in time phase. This means that the current is increasing or decreasing at all points of the conductor at the same time; also that all voltage changes at different parts of the conductor occur in phase. However, the voltage and current in the conductor are 90 degrees displaced from each other in time phase, the same as in a coil- and condenser-parallel circuit. This is shown by means of the diagrams, Fig. 179.

In position 0 on diagram (a) the current curve is at zero at the same instant that the voltage curve is in position 0 on diagram (b), which is a

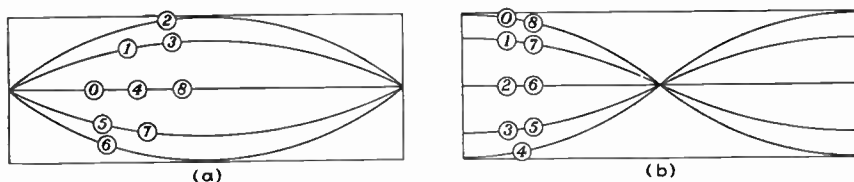


FIG. 179.—Time-phase relations of current and voltage in a half-wave doublet.

positive maximum. These figures then show the time-phase relation between voltage and current.

160. The Quarter-wave Antenna.—It has been stated before that many other radiators are developments from the half-wave doublets. This does not mean that the doublet was used first, but that one-quarter-wave antennas (T and L types, etc.) are in effect simply one-half of a half-wave-length doublet system. To demonstrate this, suppose we have a radiator with a large capacity connected to one end, such as a large sheet of copper, as shown in Fig. 180. It will be found that the current

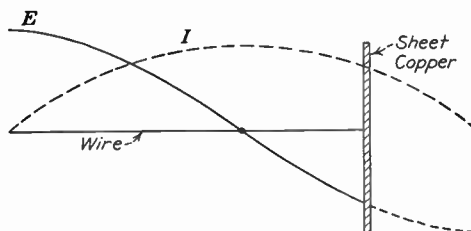


FIG. 180.—Quarter-wave antenna curves.

and voltage distribution will be changed in the right-hand section of the radiator, in that the current will not come to zero, and the voltage will not be so high as on the free end. If this copper sheet were made larger and larger, the nodal- (or zero) voltage point would gradually move closer and closer to the capacity, until finally, if the earth were connected in place of the copper sheet, and the wire brought to a vertical position, the result is a quarter-wave antenna with distribution curves as shown in Fig. 181. The frequency which gives this distribution has been called the *fundamental* of the quarter-wave antenna. For a half-wave doublet

the frequency which gives the distribution curves shown in Fig. 178 is called the *fundamental*.

If the total height of the antenna is less than one-quarter wave length, it can be loaded or tuned to resonance by means of adding inductance coils in series or between the antenna post and ground. This is sometimes necessary, although the efficiency of this system is known to be low, owing to losses in the coils and to the fact that the coil does not radiate, as would the length of antenna it replaces.

There are various forms of *flat-top* antennas which have been used for many years. One of the most common is the L type. This design consists of a wire, or several parallel wires, arranged to form an inverted L (T). The base of the L becomes the *flat top* of the antenna.

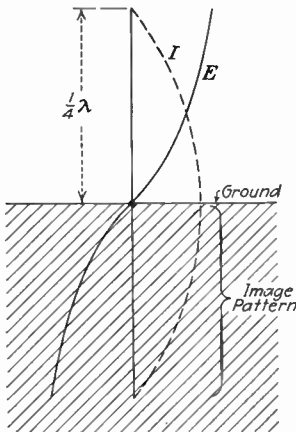


FIG. 181.—Current and voltage distribution in a quarter-wave antenna.

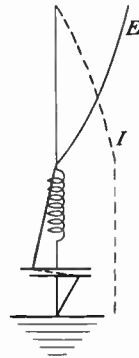


FIG. 182.—Distribution curves in a loaded vertical antenna.

The T type is similar, except that it has a flat top each side of the down lead instead of on only one side. There are various other types the names of which describes their physical shape. These are the umbrella type and the fan type, but they are no longer in common use and need not be discussed.

For many years it was common practice to operate these antennas at the *fundamental* or at a frequency slightly lower than the fundamental by means of a coupling coil inductively coupled to the transmitter-output circuit. If it was desired to operate exactly on the fundamental, a series condenser was used, its capacitive reactance neutralizing the inductive reactance added by the coupling coil. The distribution curves would then be as shown in Fig. 182. Advantage is here taken of the high ratio of radiation resistance to total resistance of the antenna, making the efficiency somewhat higher. More recently development and experi-

mentation have resulted in great progress in the design of grounded antennas. It is found that by using enough capacity reactance in the series condenser, the antenna can be operated well below the fundamental. The limit to which this can be carried is just before a half-wave is standing on the antenna. This can best be shown by a curve of the reactances involved for a vertical antenna as the height is increased and the frequency held constant. This is shown in Fig. 183.

It is understood that when tuned to resonance, the total reactance will be zero, so that, if the antenna is one-eighth wave length high (point Y on the diagram), it can be predicted from Fig. 183 that, to tune it to resonance, it will be necessary to add inductive reactance to neutralize the capacity reactance which the antenna has, as shown by the curve,

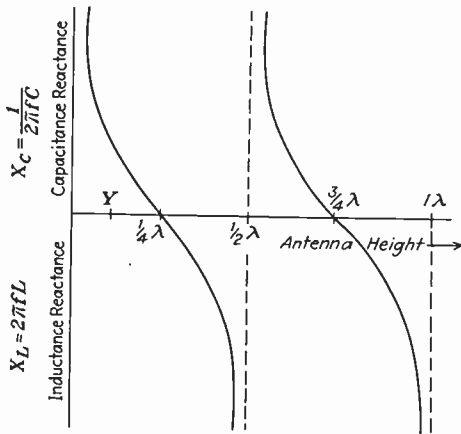


FIG. 183.—Reactance curves of a vertical antenna.

and bring the resultant to zero. How much inductance will be necessary to do this can be determined from the measurement of the capacity reactance, as described in a later section.

It should be noted that the same resonant points will also be found if the antenna height is constant and the frequency is increased. With the grounded antenna the first resonance noted will be when the wave length of the oscillator is four times the height of the antenna, in meters. As stated before, the frequency corresponding to this wave length is called the *fundamental*. The next resonance point will be when the wave length of the oscillator is three-fourths of the height of the antenna, in meters. This oscillator frequency is then called the *third harmonic* of the antenna. There are three quarter waves standing on the antenna. From Fig. 183 we can see that resonance points would be found only on the odd multiples of one quarter wave; for example, one-fourth, three-fourth, five-fourth, etc.

II. ELEMENTS OF RADIATION

161. The Radiation Field.—In the foregoing section there were discussed the voltage and current characteristic curves of the antenna itself. We shall now consider the way in which these currents and voltages in the antenna are capable of producing disturbances in the ether or medium carrying the wave motion. It is understood that the current-amplitude curves simply show the relative quantity of electrons passing a given point in a specified period of time. This electron flow produces a magnetic field around the conductor, the same as in low-frequency power circuits. In addition to the magnetic field there is also an electric or electrostatic field which is due to the potential differences of the ends of the antenna. In Fig. 184 are shown the magnetic lines of force and also the electrostatic lines which are always at right angles to them. This is sometimes called *space quadrature*.

To define what this field actually consists of we must first give some explanation of the material from which it is built; in other words, we must define *ether*. This medium is generally considered to consist of the resultant of all electric fields at the position in question. Then the electric field and magnetic fields are strained conditions of this "ether." As the current flow increases in the antenna in one direction, the magnetic and electric fields are moving out in all directions from the conductor with the velocity of light; but, as the current flow in the conductor

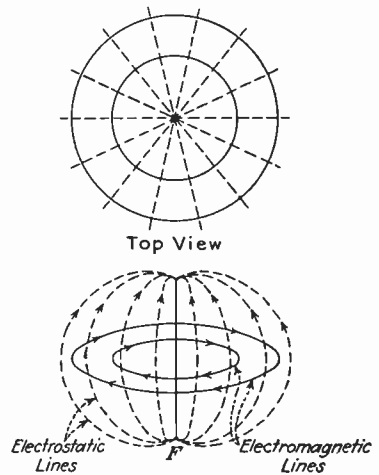


FIG. 184.—Electric and magnetic field distribution around a vertical antenna.

begins to decrease, the field or strain near to the conductor begins to collapse or move in toward the conductor. However, it takes an appreciable time for the electromagnetic field to change its direction of motion; it takes time for this new force to move out and affect the field to the full extent of the first disturbance, and in this period of time the first wave has progressed farther out. Because of this condition most of the energy is not returned to the antenna circuit as it would be in a tank circuit or in a parallel-tuned closed oscillatory circuit.

The distance from the peak of one electrostatic or magnetic disturbance to the peak of the next is called a *wave length*. The usual symbol is λ (the Greek letter lambda). Since the number of these waves which are sent out depend on the frequency or period of vibration of the oscillator or radio transmitter, it follows that the velocity of wave travel will

equal the wave length of one disturbance multiplied by the frequency of oscillation. Thus we write

$$V(\text{meters per second}) = \lambda f \quad (145)$$

In a given medium the length of a wave (radio, light, or water) will be of a fixed value, the frequency of course remaining the same. This means that if a radio wave passes through a medium other than air, it will have a different wave length and a different velocity while it is traveling in the other material.

As we mentioned above, the electrostatic and electromagnetic fields are interlinked and travel through space together; but in cross section they would be found to be 90 degrees displaced in quadrature. It can be shown that these two fields are interdependent. If energy is taken from one field, it is balanced or replaced from the other so that they are always equal in energy content.

162. Sky and Ground Waves.—The radiated wave is considered to travel by two paths toward the receiving site. The first, and for broadcast reception the most important, is the ground wave, or that part of the energy which travels along the surface of the earth. The other path, which is responsible for the success of short-wave circuits, depends on reflection from one or more of the ionized layers which are considered to exist at great heights in the atmosphere. These two propagation paths cannot be considered entirely independent, as the ground wave is reinforced by diffusion from low-angle radiation slightly above the ground wave, which is entirely different from any Heaviside layer reflection.

At night it is also found that there is a *sky wave* or reflected wave to contend with, on broadcast frequencies, as well as the ground wave, making the problem of estimating the signal strength more complex.

We shall first consider the ground wave during daylight conditions. As stated above this will eliminate consideration of Heaviside layer reflections, although there are some evidences of reflection at great distances.

It is well known that reception over salt water is better than over dry, sandy ground, the reason being, of course, that there is less attenuation of the ground wave when the path is over a good conductor. However, the ground-wave field strength can never be as great as would be indicated by the inverse-distance curve. The difference between the actual signal received and the inverse-distance curve will be due to this attenuation of the signal by the ground or water over which it passes. One of the first investigations as to the amount of this attenuation was made by L. W. Austin.¹ The formula deduced by Austin for the current which will be received during daylight over sea water at a given distance from the transmitter is as follows:

¹ U. S. Bureau of Standard Bulletin, vol. 11, p. 69, 1914.

$$I_r = AI_s \times \frac{h_s h_r}{d\lambda} \times \epsilon^{-0.0015 \frac{d}{\sqrt{\lambda}}} \quad (146)$$

where A = a constant found to be equal to 4.25.

h_r = the height of the receiving antenna.

h_s = the height of the transmitting antenna.

I_r = the effective current in the receiver antenna.

I_s = the effective current in the transmitter antenna.

Note: All lengths are expressed in kilometers.

The latter part of this equation $\epsilon^{-0.0015 \frac{d}{\sqrt{\lambda}}}$ is the attenuation factor in addition to the inverse-distance term which is (d) in the denominator of the above equation. Any formula of this type should be understood to have a limited application. There are numerous variables which will influence the results. The attenuation effect of different kinds of soil, radiation efficiency, type of antenna used, and the curvature of the earth are some of these factors.

163. Field Intensity.—The inverse-distance curve or equation will in any case be found to be the maximum signal which it is possible to receive, as it provides only for the spreading out of the wave and does not make any allowance for attenuations. There will frequently be signals of as high an intensity as this inverse-distance curve indicates, owing to reflections from the ionized layers of waves which are not attenuated by the ground; so, the equation has value in estimating the maximum signal possible in a given condition. This equation may be shown to be quite accurate for short distances where the curvature of the earth does not affect the results. The equation, when measurements are made in miles, is as follows:

$$E_{\mu v./m.} = \sqrt{P_{\text{watts}}} \frac{5.87 \times 10^3}{D_{\text{miles}}} \quad (147)$$

where D_{miles} = the distance in miles from transmitter to receiver.

$E_{\mu v./m.}$ = the received signal in microvolts per meter.

P_{watts} = the power in watts radiated by the antenna.

The field strength at a given distance from a transmitter may also be determined approximately by the ratio

$$\mu v./m. = \frac{1.25 \times I \times f \times h_{eff}}{d} \quad (148)$$

where d = the distance measured from the radiating system in kilometers.

h_{eff} = the effective height of the antenna in meters.

I = the antenna current.

The effective height (h_{eff}) of the antenna does not refer to its physical height above the ground but to its effective electric or magnetic height.

Thus,

$$h_{eff} = \frac{\mu v./m. \times d}{1.25 \times I \times f} \quad (149)$$

Field-strength measurements are generally measured in microvolts ($\mu v./m.$) or millivolts per meter ($mv./m.$) at different points from the radiating system to determine the distribution and efficiency of the field pattern. For example, let us assume that a potential difference of 100 volts is developed between the antenna system and the ground owing to excitation from the transmitter. As a result an electric field will be developed owing to the strained area between these points. Now let us assume the antenna to have a vertical height of 200 ft. or approximately 61 meters (1 ft. = 0.3048 meter). Since the potential of 100 volts developed across these two points is separated by 61 meters, the strength of the field would be 1.639 volts per meter. This is obtained by the reasoning that if a potential difference between two charged bodies is 100 volts for a distance of 1 meter, the potential difference, if the two charged bodies are moved apart to a distance of 2 meters, will be 50 volts per meter. Thus, in this example the potential difference of 100 volts is distributed in 61 meters which, therefore, means that 1.639 volts are developed for each meter. Hence it may be seen that the closer the relation of the bodies to each other, the greater will be the field strength per meter; and the farther apart the two bodies, the lower the field strength in volts per meter. Consequently, as the distance from the point of measurement to the radiation system increases, the field strength per meter decreases. The strengths of the fields, therefore, at remote points from the radiating system may decrease to such small values that they are measured in millivolts or microvolts per meter.

There have been a large number of field-strength surveys made and many of these surveys are checks on the basic formulas proposed by Sommerfeld in 1909. Using this formula P. P. Eckersley¹ has worked out a number of curves of field strength in millivolts per meter. His measurements show attenuation factors ranging from 10 to 11 c.g.s. units to 0.75×10^{-14} c.g.s. units, depending on the dampness of the ground and on its hilly or mountainous character, the mountainous country having very high attenuation.

It may be well to mention what effects are generally caused by some of the disturbing elements to propagation which were mentioned above. It is thought that a sun spot or a flash of ultraviolet light from the sun has an effect on the magnetic activity on the earth, and, owing to this disturbance or to the effect of the ultraviolet radiation on the ionized Heaviside layer, the propagation of radio waves is affected. Nevertheless, it is very noticeable on short-wave radio circuits that, when there

¹ *Proc. I.R.E.*, July, 1930, p. 1170.

are aurora borealis displays, there is a pronounced drop in signal strength on east-west circuits, especially those operating in the northern sections. The effect is not nearly so pronounced on north-south circuits. The effect does not seem so pronounced on broadcast frequencies; and on long waves there is sometimes an increase of signal reported in the daytime. On short waves below 75 meters, this effect is so pronounced that many short-wave circuits become entirely inoperative during these magnetic storms. It is said that one of the leading communication companies in this country keeps its old long-wave transmitters very busy during these periods.

Other losses which are considered to have some effect on radio waves are eddy currents and magnetic hysteresis. There has been noticed pronounced weakening of broadcast signals measured in a test car on the far side of a downtown New York section. This seemed without doubt due to absorption or screening by the tall steel buildings in the propagation path. Dielectric losses are often due to trees, buildings, or wooden masts in the field of the antenna, especially when the mast is not dry and painted. Ionization of the air probably has an attenuating effect. Thus it is considered that the excessive ionization during daylight is responsible for the lack of Heaviside layer reflection at broadcast frequencies during the day.

At night the conditions are entirely different. The ground wave is present the same as before, but in addition there is a reflected wave at broadcast frequencies which generally has an amplitude comparable to that of the ground wave at a distance which varies from about 35 to 60 or 70 miles depending on which end of the broadcast-frequency spectrum is involved, the antenna design, and the conductivity of the ground. At this circle around the transmitter site there will be pronounced fading at times, owing to the fact that the reflected ray will be out of phase with the ground wave, and the fields from the two sources will cancel. Because of this fading and poor reception, at points near and beyond this distance, it is becoming common practice to estimate the service area of a broadcast station as the area well within this circle. It can be seen that no increase in power will change the position of this *fading circle* because both the sky wave and ground wave are increased in intensity in the same degree. The amount of power to use is then determined by the signal strength necessary within this area. This, however, may be as high as 50 kw. or more. It can be seen that the lower frequencies of the broadcast band have a pronounced advantage from this standpoint, since the distance at which fading will start, will be at a greater distance from the transmitter. This means that a proportionately greater *service area* will be found for low-frequency stations. The signal beyond the service area depends almost entirely on the sky wave and will fade at times, the same as short-wave signals do, because of changes in the Heaviside layer. However,

it may often be considered as a fairly good signal by the average listener when stronger and more constant signals are not present.

164. Field-strength Measurements.¹

One of the simplest methods for measuring the field strength of a radio wave is that illustrated in Fig. 185. The loop antenna forms the pick-up circuit for delivering the signal energy to a sensitive receiver preferably of the superheterodyne type. The condenser and loop are adjusted to the proper LC value to be resonant at the desired frequency to be measured. A milliammeter is connected in the output-plate circuit to indicate conditions of resonance with input-signal frequency.

After the loop circuit has been adjusted to resonance with the incoming signal and the reading noted on the output milliammeter on some convenient point on its scale, the degree of deflection being regulated by the receiver volume control or a separate regulator as illustrated, the loop is slowly turned to a right-angular position to the plane of the oncoming wave until the meter reads zero deflection. The local oscillator-amplifier is then set into operation in the usual manner and the output circuit adjusted to maximum deflection to resonance as indicated by the radio-frequency ammeter M_2 . If a potential divider resistance R_1 , of proper wattage and of non-inductive design characteristics, is connected across the output of the signal generator a radio-frequency current will be forced to circulate through it of a value $I = E/R_1$.

The voltage across R_1 may then be used to excite the loop during the

¹ A thorough analysis of field-strength measuring equipment is given by C. R. Englund

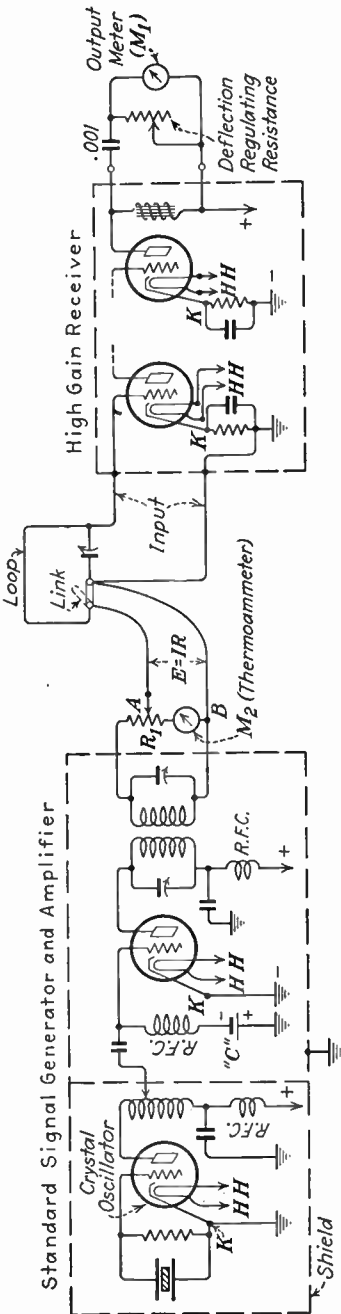


Fig. 185.—Field-strength measuring circuits.

and H. T. Friis, *Trans. A.I.E.E.*, vol. 46, p. 492, 1927.

period in which it is at right angle to the plane of the oncoming wave so that direct comparisons between the signal and local generator may be made. The potentiometer is adjusted so that the voltage E_0 is of such a value that, when the loop link switch is open, the output milliammeter M_1 will give the same reading as when the loop had been adjusted to resonance with the incoming signal. When the two signals produce identical readings on the output meter, the field strength of the incoming wave can be measured, according to Terman, by the ratio

$$\epsilon = \frac{E_0}{h_{eff}} \text{ volts per meter} \quad (150)$$

where E_0 is the tapped voltage developed across R_1 , and h_{eff} is the effective height of the loop antenna per meter.

The effective height of a loop antenna may be very readily calculated by the formula

$$h_{eff} = \frac{6.28 \times l \times h \times n}{\lambda} \text{ meters} \quad (151)$$

where l = horizontal length of the loop in meters.

h = vertical height of the loop in meters.

n = number of turns on the loop.

For example, let us assume the loop to have a horizontal length of 3 meters, a vertical length of 3 meters, and 10 complete turns. The loop is tuned with a small variable-shunt capacity to a wave length of, say, 400 meters. Thus

$$h_{eff} = \frac{6.28 \times 3 \times 3 \times 10}{400} = 1.3 \text{ meters (approximate)}$$

Now if the voltage developed across the output of the local generator at E_0 is 2 volts, the field strength will be

$$\epsilon = \frac{E_0}{h_{eff}} = \frac{2}{1.3} = 1.53 \text{ volts per meter}$$

The voltage E_0 is computed by the equation $E_0 = IR_1$, where I is the current flow through M_2 , and R_1 is the amount of resistance in use between the points A and B of the voltage divider under conditions when the output meter M_1 reads the same deflection as when the incoming signal was received.

165. Propagation of Short Waves.—As has been suggested above, the propagation of short waves is much more involved owing to reflections from the sky wave both during full daylight and at night. In this discussion of short waves we shall consider wave lengths below 60 or 70 meters. At these frequencies, it is found that the ground wave is very rapidly attenuated, so that the reflected wave is depended on entirely for communication.

The Kennelly-Heaviside layer or layers are considered to consist of air and gases which are ionized by sunlight, and ultraviolet radiations from the sun. This ionization process is in a continual state of change, more electrons are breaking away from their molecules, and other electrons are recombining with positive ions to form molecules again, there being a much higher degree of ionization during the day and a gradual change in the evening and early-morning transition periods. The height of these layers and the extent of their ionization will determine the absorption, refraction, or reflection of the wave.

There have been numerous measurements made by various laboratories with results which indicate very strongly that there are at least two reflecting layers. The lower layer is now being called the *E* layer and the second the *F* layer. J. P. Schafer and W. M. Goodall,¹ have found *virtual heights* for the *E* layer ranging from 90 to 125 km., and for the *F* layer heights between 240 and 350 km. with *critical-period* heights as high as 650 to 700 km. These critical periods are the time at which there is a transition from one layer to the other and occur in the evening or early morning.

From these heights and the density of ionization it is possible to estimate skip distances, and the probable amount of refraction and absorption. It should be understood that a wave may be reflected from the lower *E* layer, or it may pass through the *E* layer, being attenuated and refracted and then passing on to the *F* layer, where it would be reflected downward again, and would again pass through the *E* layer being attenuated and refracted a second time; it would then continue on to the surface of the earth. Whether the wave is reflected from the first layer or whether it penetrates the first layer and goes on to the second, is determined by the frequency of the wave and the extent of the ionization.

The reason that these ionized layers reflect might be considered to be due to their conductivity. The effect is somewhat similar to placing a copper sheet over the path of the wave. Of course, the ionized layer could not be homogeneous and concentrated as a copper sheet, and this accounts for the variations in intensity of the signals which are reflected from these layers.

Another possibility is that the wave is not actually reflected at all but is refracted more and more, until it is bent back toward the earth. This is shown in Fig. 186. In either case there may be considered as existing a *critical angle* which for a given frequency is the smallest angle of incidence from which the wave will be returned to the ground. This angle φ is shown in Fig. 186. Any radiation which strikes the layers at less than this angle (φ) will penetrate them and be absorbed.

¹ *Proc. I.R.E.*, July, 1932, p. 1131.

166. Skip Distance.—The first point at which a reflected wave reaches the surface of the earth would be at point *S*, Fig. 186. Since the ground wave travels only a short distance such as *TG*, it can be seen that the distance from *G* to *S* will have no signal. The limit to the low-angle radiation is the lowest which will be propagated without absorption by buildings, trees, and other obstructions. Then the area from *S* to *S'* is completely covered by the reflected waves. In general, it is found that the critical angle α may be smaller for low frequencies without decreasing reflection from the ionized layers. This means that a lower frequency should be used for short- than for long-distance transmissions.

It is naturally to be expected, and is approximately true, that the greatest ionization occurs at noon when the ultraviolet rays reaching the earth are most intense. At this time, the reflecting layers, which

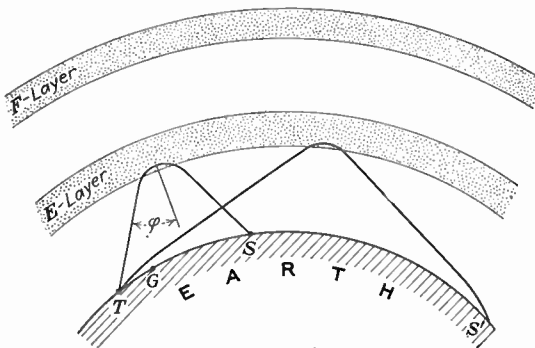


FIG. 186.—Reflecting layers and signal paths.

consist of ultraviolet radiations, are nearer the earth, resulting in a shortening of the skip distance *GS*. Recall again now the fact that, in general, a small critical angle is characteristic of a low frequency and a large critical angle is characteristic of a high frequency. It is seen then that a higher frequency can be used in the daytime, when the layer is low, than at night, when the layer is high, without affecting the reflection of the wave to the earth. With a given angle, it is seen that, as the reflecting layer is raised, the skip distance increases. Hence, if a high-angle, high-frequency wave is used at night, the reflected wave will pass far beyond the earth owing to the greater height of the reflecting layers. By keeping the angles shown in Fig. 186 fixed, it is easily seen how the skip distance varies with a change in the height of the reflecting layer or layers. Average experimental values for the skip distance found in summer and winter for day and night transmissions at two frequencies near the opposite ends of the high-frequency spectrum are given in the following table:

TABLE IIIB.—APPROXIMATE SKIP DISTANCES

Frequency kilocycles	Approximate skip distance, miles			
	Summer		Winter	
	Day	Night	Day	Night
6,000	90	300	150	750
15,000	500	2,800	1,000	10,000

In general, then, the best frequency to use is the one which gives the desired reflection. Therefore, if the distance between the transmitter and receiver is known, and the frequency which has a skip distance up to

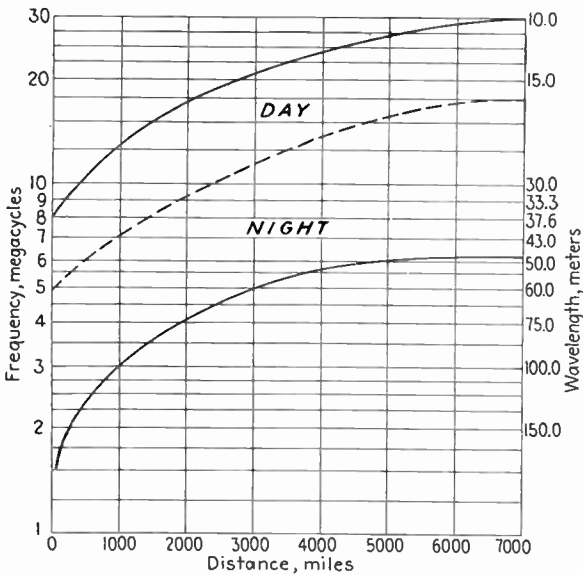


FIG. 187.—Skip-distance curves.

this point is also known, it is only necessary to decrease this frequency about 20 per cent to obtain the optimum frequency. This is an approximate method. Some comprehensive charts on propagation over the north Atlantic Ocean have been prepared.¹ Another paper² shows propagation characteristics in all directions from Schenectady, N. Y., and are very valuable for anyone choosing frequencies for a radio circuit. It must not be assumed, however, that from year to year conditions will stay the same, as there have been pronounced variations in certain years, probably due to sun spots or magnetic storms. A chart showing the

¹ BURROWS, C. R., *Proc. I.R.E.*, September, 1931, p. 1634.

² PRESCOTT, M. L., *Proc. I.R.E.*, November, 1930, p. 1797.

approximate relation of optimum frequency to distance in short-wave transmission is shown in Fig. 187.

167. Echo Signals and Multiple Reflections.—The method used by the U. S. Bureau of Standards and several other laboratories in determining the height of ionized layers has been to key the transmitter with a short, sharp pulse, and to record the pulse as received from the ground wave, and again from the reflected sky wave which will arrive at a later time since it travels a greater distance. From this time, as shown by the distance on the oscillograph-photograph film, the height of the layer is calculated. The term *echo* has been applied to these reflected signals which arrive later than the ground wave. It is entirely possible for echo signals to distort the received signal when transmitting at high speeds, or when the signal is modulated.

Multiple reflections probably account for many cases of short-wave reception at very great distances. It is thought that, after returning to earth from the first heaviside layer reflection, the wave is reflected from the earth back to the layer, again and again.

168. Propagation of Ultra-short Waves.—Ultra-high-frequency radiations are quasi-optical in nature; that is, their line of propagation is almost a straight line. The signals attenuate rapidly when the straight line between the transmitter and receiver becomes tangent to the earth's surface. It is possible, therefore, to increase the communication range by elevation of either the transmitter or receiver.

Roughly speaking, the ultra-high-frequency spectrum between 5 and 10 meters is 30 times as wide as the broadcast band. In other words, by spacing stations 10 kilocycles apart it is possible to squeeze in about 100 stations in the broadcast band; while, with the same separation, 3,000 stations could operate in the 5- to 10-meter band. Therefore, when we speak of the ultra-high-frequency band, it is necessary to specify which end of the band we mean. The conditions on 10 meters, for example, might be vastly different from those encountered on 3 meters. This is true of wave attenuation on these ultra-high frequencies. As the period of oscillation approaches the infra-red spectrum, the optical similarity between radio waves and light waves becomes more apparent.

For wave lengths not longer than 7 meters the wave passes right through the Kennelly-Heaviside layer, and there is no sky wave reflected back to earth. This leaves only the immediate radiation from the antenna and the energy in the ground wave to affect the receiver. The range is, therefore, essentially limited, and it becomes possible to operate a great many stations on the same frequency.

Owing to the quasi-optical nature of these ultra-short waves, pronounced radio shadows are noticeable. Mountains, forests, and tall buildings all cause the signal to be sharply attenuated or to disappear altogether when these obstructions come between the transmitter and the

receiver. The extent of the signal attenuation depends on the frequency used. At the extremely high frequencies even tree trunks cause a radio shadow.

Radio shadows are caused by the inability of the ultra-high-frequency waves to bend around obstructions, and this effect becomes more pronounced as the infra-red spectrum is approached.

There is an absence of static on the ultra-high-frequency spectrum, and communication can be carried on during severe electrical storms. Interference from man-made static, such as from automobile-ignition systems which are not shielded, may become serious, but this interference can be eliminated entirely by proper shielding of the automobile-ignition systems and other generators of spark discharges.

Owing to the limited range of the signals the service area of a given station can be very accurately predetermined and limited. This naturally reduces interference between stations, although they may be located relatively close together.

169. Polarization.—Polarization as applied to radio waves may be defined as the angle at which the greatest field is produced. As an example, suppose that a vertical radiator is being excited by a transmitter, and a short doublet with an indicating meter at its center is held within two or three wave lengths of the vertical antenna. It will be found that the greatest current will be induced in the doublet when it is perpendicular to the ground. Hence, the wave is still vertically polarized. Now suppose that the doublet is taken a greater distance away and a receiver is substituted for the less sensitive indicator. The results then are likely to show a greater voltage received when the doublet is turned one way or another from the vertical, and it is not uncommon to find (at high frequencies) the greatest signal on a horizontal receiving doublet, even when the transmitting antenna is vertically polarized. This is often found at frequencies above 3,000 kilocycles. At long wave lengths and at broadcast frequencies, this twisting of the polarization of the wave is not common.

The exact cause of this twisting of the polarization of the wave is not known. It may accompany the reflection or refraction process in the Heaviside layer. Many measurements have been made at high frequencies of the polarization of waves from both vertical and horizontal radiators, but results vary. Sometimes a vertically polarized wave predominates at the receiver, and at other times a horizontal wave is stronger. In general, slightly better signals are received on a horizontal antenna at high frequencies than on a vertical wire. It is also more easily constructed.

In an earlier section were explained and shown the current- and voltage-distribution curves of a half-wave doublet or dipole and a quarter-wave antenna. These curves do not in themselves show the combined

field intensities around the antenna, but it is possible to determine mathematically what the distribution or intensity of field will be in any direction if the current and voltage distribution and amplitudes are known; or it is also possible to determine the field experimentally. This is done by measuring and plotting graphically the intensities of the radiation from all angles of an antenna, and this is accomplished by means of small pick-up loops, or a short doublet, and a sensitive thermoammeter or vacuum-tube voltmeter at close ranges. When it is desired to make field measurements at distances of several miles, a calibrated receiver or a receiver and a standard-signal generator with a calibrated attenuator are used.

The field which is found to exist around a half-wave doublet as described above is in the form of a torus or ring around the conductor at its center. A cross section of this field and the antenna conductor is shown in Fig. 188. This is the field of a vertical half-wave radiator. If the dipole were turned so that the conductor is horizontal, the field would stay in the same position relative to the conductor. This field distribution could be expected because of the fact that the current is greatest at the center of the wire.

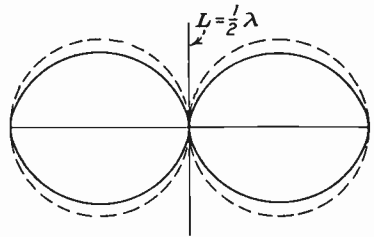


FIG. 188.—Field pattern of half-wave doublet.

The maximum signal would then be received on the broadside or at right angles to the antenna if the doublet is horizontal; but, if the doublet is vertical, there is equal directivity in all directions in the horizontal plane. There is no radiation along the length of the wire. It can be shown that the cross sections are not true circles but are slightly compressed in the vertical plane. If the capacity of the ground is taking the place of the lower half of the radiator (a one-fourth wave vertical), the resultant field is found to be simply the upper half of the torus ring. It is non-directional in the horizontal plane. Even if the quarter-wave antenna has a flat top or capacity added to its top, it will not be very directional unless the flat top is long compared to the vertical height. If the flat top is long and has large capacity to ground, the horizontal projection of the radiation curve will no longer be a circle but will be elongated in the same direction as the length of the flat top and will be slightly more directional in the direction of the lead in.

III. RADIATING SYSTEMS

170. Short-wave Antenna Design.—It would be expected that a half-wave doublet would be exactly one-half wave long; however this is not exactly true, in practice, because the antenna has to be held up by insulators and the effect of this rigging is found to add to the actual length

of the radiator; so it is necessary to use a correction factor to determine the actual length to make the conductor of the doublet. If insulators are used which are long and slender and do not have metal-end caps, the curve of Fig. 189 may be used to determine the approximate correction factor for the above *end effect*. This curve has been determined experimentally but does not fit all cases. The procedure for best results is to tune the half-wave doublet exactly to the transmitter frequency, starting with the corrected length determined from the above mentioned curve.

There are several methods of determining when the doublet is exactly in resonance. A thermal radio-frequency meter may be inserted in the center of the doublet or shunted across a short section of the conductor at the center and read by means of field glasses. A radio-frequency

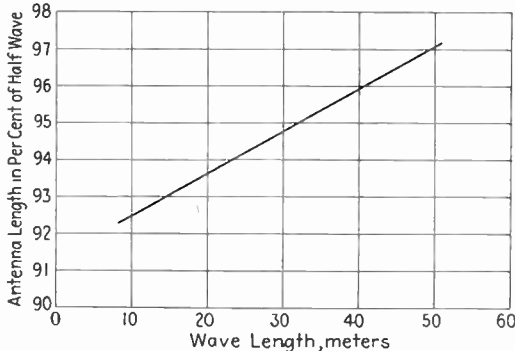


FIG. 189.—Curve giving end-effect constant in percentage.

meter and a pick-up loop can also be used coupled to the line feeding the antenna. This latter method will be described under Transmission-line Feed Systems.

Another method is to erect a pick-up loop or short doublet with indicator, as described above, about three wave lengths in front of antenna, and then to tune the antenna until the greatest signal strength is secured. Very little electrical difference is found when using different-sized wires for the radiator, a No. 8 or 10 size will be as good as a larger wire; but mechanically it is important to use a heavy enough wire to withstand the extra strain incurred from sleet formations which sometimes occur, or where the span is great.

The question of height is an important factor in relation to the cost of a transmitting antenna for short-wave work. There are many different types of directional short-wave antenna designs, and some require tower heights of 300 ft. or more, making a very expensive design; but, for an ordinary horizontal doublet, there will be only a small gain found by increasing the height above one wave length. This generally means that wooden poles can be used which also necessitates the use of guys. It is much more important to have guys and halyards broken up by insulators when the antenna is working at high frequencies than when it

is working at the lower broadcast frequencies, because even a short guy can act as a reflector for the antenna and distort its directivity pattern as well as absorb power. If the egg type of strain insulators is used, it is necessary to place them closer together because of the high capacity between the ends of the wire at the insulator. Egg insulators are generally used, however, on heavy poles because of their high compression strength. A spacing of approximately 12 ft. between eggs should be sufficient for frequencies as high as 15,000 or 20,000 kilocycles.

171. Broadside Directional Antenna.—Most short-wave directive-antenna designs are found to be different combinations of half-wave

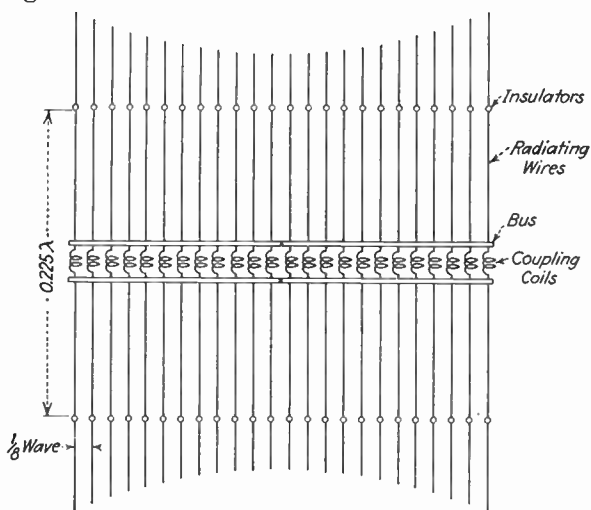


FIG. 190.—Broadside directional antenna.

doublets, arranged so that the radiations will add in one direction. However, the RCA Communications' *broadside directional antenna* is an exception to this rule. This antenna consists of two rows of vertical radiators, each radiator of which is less than a quarter wave length long. The spacing and length of one bay of the antenna are shown in Fig. 190. The back row is five-fourths wave length in back of the front row and is generally used as a reflector for the front row, but this reflector is not of the *floating* type. It is not excited by radiation from the antenna itself but from the transmitter, being coupled into the transmission line which feeds the antenna. It is not fed in phase with the antenna, but with a leading or lagging phase angle depending on whether it is desired to transmit front or back. The directivity is at right angles to the rows of radiators or *broadside* as the name implies.

The method of feeding the rows of radiators is by two copper bus bars (in some antennas four bus bars are used in an arrangement to melt sleet) at the center of the radiators with a coil across the buses for coupling. The capacity of the buses can be adjusted by changing spacing, so that an effect on the bus feeder of a velocity greater than light is secured.

The object of this is to keep the excitation of all doublets in phase. However, this can only be carried to certain limits, and each bay of these antennas contains only 24 radiators spaced one-eighth wave apart. If greater directivity is necessary than is secured in one bay, more bays are added, each bay consisting of a row of radiators and a row of reflectors and each bay being fed at its center.

This antenna was one of the first short-wave directive antenna designs used commercially. However, the RCA has developed other types which will probably come into greater use in the future. One of these later designs is called the V type and is based on radiation from a long wire operating with several waves standing on it.

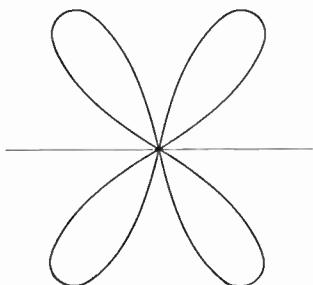


FIG. 191.—Field pattern, one wire 1λ long.

Figure 188 showed the field around a half-wave doublet; but the number of ears and their distribution with respect to the wire will be found to change depending on the number of half-waves standing on the wire. If the wire is one wave length long, the field intensity in cross section will be found to be as shown in Fig. 191. Each additional half-wave added to the length of the wire will add two more ears to the above given cross-section view.¹

It has been shown that these ears are not all of the same size or intensity, and that two of these long wires can be arranged in the form of a V at such an angle that two of the ears from one wire will be superimposed on two of the ears of the other wire. This is shown for two wires, each one wave length long, in Fig. 192. The two wires may be joined at the center of the V as shown.

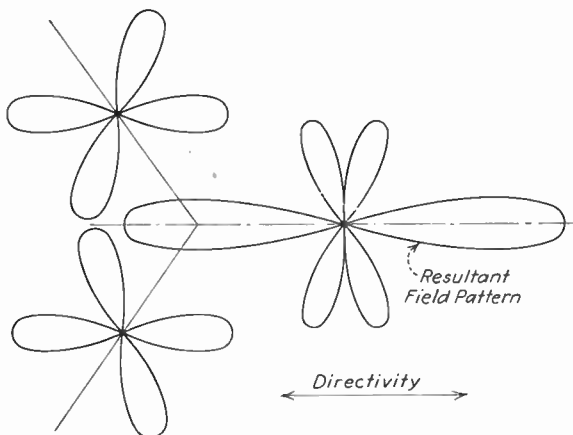


FIG. 192.—Field pattern, two wires each 1λ long.

Excitation can be accomplished by tapping the transmission line across a small portion of the radiator near the point of the V, which is the same method used in a doublet antenna. It must be remembered that the wires would then have to be of such a length as to make a current-maximum point come at the point of the V and a voltage peak at the ends of the V. This would mean an odd number of quarter waves

¹ CARTER, HANSELL and LINDENBLAD, *Proc. I.R.E.*, October, 1931, p. 1797.

on each wire. To accomplish this the wires would have to be tuned by changing their length, which means lowering the antenna.

An easier method to accomplish this is to use approximately an even number of quarter-waves of length in each wire; not joining the wires, but letting an additional quarter-wave of wire on the end of each extend down the supporting pole at the point of the V. The exact tuning is then accomplished by means of a shorting bar across these two ends, arranged so that it can be moved along the two wires. This method is described more in detail in a later paragraph.

The commercial form of this antenna has two V radiators, one above the other on the same three poles, and another identical system as a reflector spaced $2\frac{1}{4}$ wave lengths in back of the first. It is excited in leading or lagging phase with respect to the antenna, depending on whether it is desired to transmit front or back, similar to the broadside reflector before described.

Both of these types of antenna will be directional in the desired direction and also in the reverse direction, unless these reflector systems are used.

These antennas have a low radiation angle in the vertical plane, which is highly advantageous as was shown in Secs. 166 to 168.

172. The Franklin Antenna.—An antenna developed by C. S. Franklin for the British Marconi Company has been used at several commercial stations in the United States, for both receiving and transmitting; but the cost of construction makes it practically prohibitive. The theory of its operation will be described, however.

As was shown in Fig. 191 radiation from a wire several half-waves long will not be at right angles to the wire but will have several ears. If it were possible to cause the current distribution to take the form shown in Fig. 193, the radiations from each half-wave section would add, and the total radiation would be at right angles to the wire. The method of accomplishing this distribution is to place a *phasing coil* between each half-wave section. The current in the phasing coil is then the reverse phase of that in the half-wave radiators; but the coil does not radiate appreciably, so there is no detrimental effect to the radiation pattern and all the half-wave radiators have the same current and voltage phase. Instead of an inductance coil, a half-wave of wire is actually used, but it is arranged in loops which cancel all radiation.

The height of the towers used for this antenna in the United States is approximately 300 to 550 ft., which allows several half-doublers end to end with phasing coils, as described above. Approximately 16 to 24 radiators are hung from a halyard between the towers, and behind the radiators, at a distance equaling an odd number of quarter-waves, and a reflector net is arranged somewhat like the broadside antenna, but, of course, of much greater height. The radiators are fed at the bottom through impedance-matching networks.

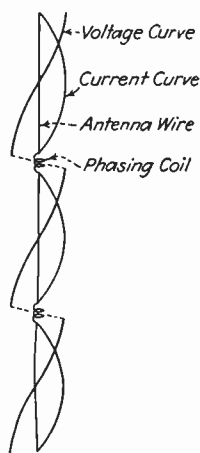


FIG. 193.—Effect of phasing coil on radiations from a series of half-wave doublets.

One of the simplest directive designs is shown in Fig. 194. The proper phase is secured for these radiators by coupling by means of half-wave doublets, but the *coupling doublets* are prevented from radiating by having two of them, which are 180 degrees out of phase, spaced close together, so that they do not radiate. This is shown by the current curves on the wires.

173. The diamond-type antenna was developed for receiving purposes by the Bell Laboratories and appears to be very economical as to first cost and is very effective. As the name implies, it is arranged in a diamond shape. The principle of operation seems to be practically the same as the V type described above. In fact, if two V types are arranged with the open side of the two V's together, the result is a diamond

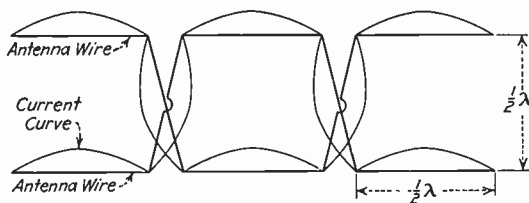


FIG. 194.—Simple directive antenna.

antenna. This similarity holds down to the angle between the wires. In the V type described above, the transmission line is worked into the antenna, but the impedance is matched in either direction. In other words, the antenna can be used to receive just as well as to transmit.

The diamond type, of course, has two points at which the impedance must be matched; at the transmission line, the same circuit can be used as is used with a V type; but at the opposite "point" the wires must work into a resistance which is equal to their surge impedance. This is to prevent reflections on the antenna and, in effect, makes it non-directional. The diamond type does not seem to be so well adapted to transmitting because of the difficulty in making for it a non-inductive termination resistance large enough to dissipate sufficient power. However, it is highly efficient for receiving.

174. Feeding the Antenna.—In several of the antenna designs previously described, a transmission line is used to couple the transmitter to the antenna. The operation and design of these lines will now be considered. In Fig. 195 is shown a coil connected to the ends of two closely spaced parallel wires, and it is assumed that a radio-frequency current is induced in the coil from a transmitter or an oscillator. If the two ends of the wires are left open at the right side of the figure, there will result very pronounced standing waves on the wires, similar to those on an antenna (there will not be as much radiation, however, because of the narrow spacing of the wires). There will also be standing waves if the

two ends of the wire are connected together, but, if a non-inductive resistance equal to the *surge impedance* of the line is connected across the end of the line, it will be found that the standing waves will have disappeared entirely.

Standing waves are generally referred to as electrical or magnetic reflections from the point of termination in a line back to the source of excitation. This condition is prevalent in any transmission line in which the load or terminating impedance of the line is improperly matched. The impedance introduced into a line to prevent reflections or standing waves is generally known as the *characteristic impedance* or *surge impedance*.

This surge impedance (Z_o) under these conditions is given mathematically as

$$Z_o = \sqrt{\frac{L}{C}} \text{ ohms} \quad (152)$$

where L = inductance of unit length of the line.

C = capacity between wires.

Physically Z_o may be considered as a resistance which replaces an infinite extension of the length of the "line" to the right of Fig. 195.

It can be seen that a half-wave dipole radiator when properly tuned represents a pure resistance (no inductive or capacitive reactance) if measured across any two points at equal distances out from the center. This resistance would increase until at the two ends of the radiator it would be several thousand ohms. Since the transmission line described above must work into a pure resistance to prevent reflections, it can be seen that a dipole or doublet can be substituted for this resistance, if the line were tapped on the dipole at the right distance out from the center, to give the correct matching resistance. This system is found to be far superior in efficiency to any other coupling system, if the distances are appreciable. These lines are run in commercial short-wave circuits for several thousand feet. It is common to obtain an efficiency as high as 90 per cent on a line 1,000 ft. long.

A method of adjusting the antenna coupling to match the impedance of the line will be described in Secs. 176-177. It should be mentioned that the two-wire open line is only one type. There are also three-wire lines (the center wire being grounded), four-wire lines, and concentric-tube lines in commercial use. In the four-wire lines, the two opposite wires are generally connected together at the ends, and the line then becomes a two-wire circuit. The advantage is an immunity to stray pick-up on the transmission lines and this type is especially applicable to receiving purposes.

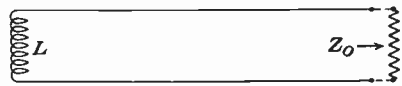


FIG. 195.—Impedance Z_o connected across a transmission line.

The concentric-tube line generally consists of two copper tubes, or a wire through a copper tube. The two conductors are separated by porcelain- or glass-insulator rings held in place on the inside of the outer tube by slight crinkles in the surface of the copper. The outside copper tube is generally grounded and, as the capacity to ground of the two conductors is not the same, it is not balanced to ground. This must be considered in designing the coupling circuits. The efficiency is generally not quite so high as the open two-wire line.

The impedance of any of these types of lines depends on the area of the conductor and the spacing between conductors. For a two-wire line the impedance is easily calculated from the equation

$$Z_0 = 276 \log_{10} \left(\frac{D}{d} + \sqrt{\frac{D^2}{d^2} - 1} \right) \quad (153)$$

This equation may be simplified, since in practice the term

$$\left(\frac{D}{d} + \sqrt{\frac{D^2}{d^2} - 1} \right)$$

is equal to $(2D/d)$. Substituting this in the above equation,

$$Z_0 = 276 \log_{10} \times \frac{2 \times D}{d} \quad (154)$$

where Z_0 = surge impedance or characteristic impedance of the line.

D = space between conductors in inches (center to center).

d = diameter of conductors in inches.

For a concentric-tube line, the equation for surge impedance is

$$Z_0 = 138 \log_{10} \times \frac{D}{d} \quad (155)$$

where D = inside diameter of outside tube in inches.

d = outside diameter of inside conductor in inches.

The method of terminating the line, as described above, for a doublet also applies to several of the other types of antenna, such as the V and diamond types.

The transmission-line feed system is the most common method of exciting the antenna in commercial use today, but there are other circuits which have some application. It can be seen that a doublet excited by the type of line described above cannot be used at a harmonic, because the percentage of coupling would then be wrong, even if a harmonic were chosen which gave the correct phase relation. However, there are certain voltage- and current-feed arrangements which make operation on several frequencies possible. These feed methods are adaptations of the Lecher-wire system which will be described before proceeding further.

Figure 196 shows the voltage and current distribution on the Lecher wires, but, if one curve is used to show the voltage, it becomes the voltage between wires and not to ground as with the doublet in Fig. 178. The Lecher wires may be shown to be themselves a development of the

half-wave dipole. The voltage- and current-distribution curves of a dipole will remain practically the same if the ends of the wires are bent around parallel. Now take two half-wave dipoles and bend the two ends of one to the right and the two ends of the other to the left, as shown in Fig. 197. The ends can be joined, since the ends of each doublet are high-impedance points. The result is then a Lecher-wire system, and the distance between current peaks is one-half wave length.

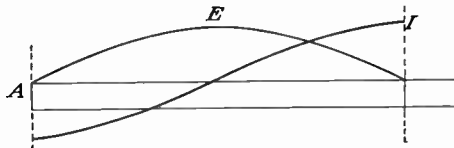


FIG. 196.—Voltage and current distribution on Lecher wires.

The original use of the Lecher wires was to determine the frequency of an oscillator. This is done by measuring the length of the standing wave (distance between current peaks) on the wire with a meter stick or scale, the frequency then being calculated from the formula

$$f = \frac{v}{\lambda} \tag{156}$$

where $\lambda = 2 \times$ the above measured value in meters.

$$v = 300,000,000.$$

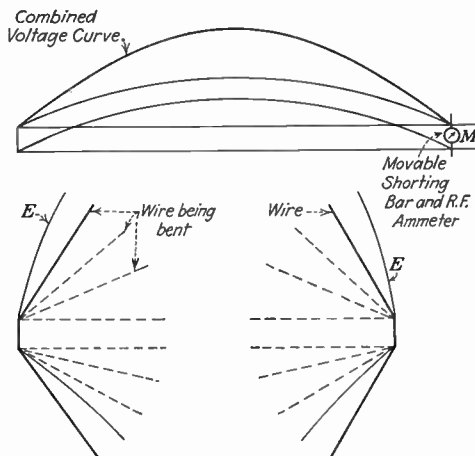


FIG. 197.—Half-wave dipoles bent around to form the equivalent of a Lecher-wire system.

A shorting bar which may be moved along the pair of wires is provided for varying the natural period of the system.

This is still a convenient and accurate way to make measurements at high frequencies. A thermoammeter is used in series with the shorting bar to indicate resonance as the bar slides along the parallel wires. The oscillator which is being measured should be as loosely coupled to the parallel wires as possible while still getting a sufficient reading.

175. Transmission Line as a Phase or Voltage Transformer.—The Lecher-wire circuit may be considered to be a transformer, because it is possible to obtain large voltage changes of the output voltage compared to the input. Suppose the termination of a transmission line is made as shown in Fig. 198. The distance to the connection points B, B' , to the right of A , is adjusted to match the impedance of the line and is found experimentally. Then if another line of higher impedance is to be coupled to this "transformer," it can be connected at a still greater distance to the right of A .

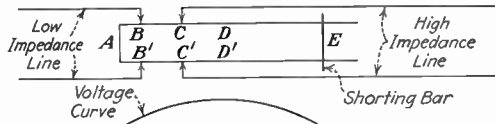


FIG. 198.—The transmission line as a phase or voltage transformer.

If the positions C, C' are moved to the right of D, D' it can be seen from the voltage curve that the impedance will decrease again. Since there is a standing wave on the wires, the current is in the same time phase in all parts of the system from A to E . To obtain a phase shift, it is only necessary to use a line with its impedance matched. In this case a distance one-quarter wave down the line will have 90 degrees time phase with the starting position. This is the method used to excite reflectors of directional antennas.

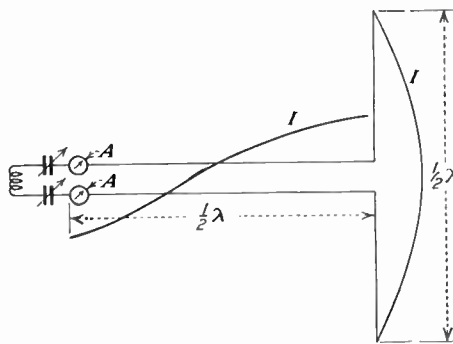


FIG. 199.—Current-feed system.

If an insulator is placed in the center of a half-wave doublet, the impedance across the insulator will be very low, since it is at a current peak. This low impedance corresponds to that across the shorting bar of a half-wave Lecher-wire system. In Fig. 199 this current-feed system is shown with an additional transmitter-coupling circuit on the left. The two condensers with the coil tune to series resonance, so that the combination has zero reactance. Practically, if the line is not exactly one-half wave length long, the error can be compensated for by the condenser tuning.

Voltage-feed systems must, of course, couple to a high-voltage point on the antenna, which on a half-wave doublet will be either end. The curves (Fig. 196) show a voltage peak at a distance one-quarter wave from point *A*. It was stated that the above voltage curves represented the difference of potential between the wires. If the curves of potential of each wire with respect to the nodal point are drawn, it will take the form of Fig. 200. If one wire is extended one-half wave length, it will radiate because there is no wire close to it of opposite polarity. It has then become an antenna.

Many low-power transmitters are directly coupled to the transmitter, but for wave lengths below 100 meters this is generally not the best practice because it means that the antenna must be close to buildings, which greatly increases losses, and lowers the effective height. Probably

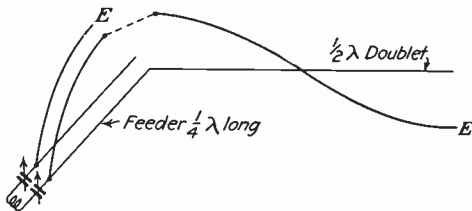


FIG. 200.—Voltage-feed system.

the commonest type of direct coupling is effected by means of a series-antenna-coil coupling to the tank circuit of the output stage of the transmitter, with a series condenser for tuning to resonance. This circuit is shown in Fig. 133.

Comparing the above systems of antenna excitation, it may be stated that the transmission line with a matched surge impedance does not have radiation losses as do the *standing-wave feeders*; also the length of the matched-impedance line can be any length which is convenient. The current-feed system must be some multiple of a half-wave long, and the voltage-feed system must be an odd multiple of a quarter-wave long. This may be some advantage with regard to operation at harmonics of the antenna. The current-feed system described above may be used at odd harmonics, as the third or fifth, and the voltage-feed system may be used on a second harmonic by changing the coupling coil to a parallel-resonant circuit. However, the field patterns will change, as was shown in Figs. 191 and 192, for a long wire.

176. Broadcast Transmission-line Coupling Circuits.—In high-power installations it is often found desirable to keep the transmitter building out of the more intense radio-frequency field of the antenna. This is accomplished by the use of a transmission line similar to the matched-impedance lines, before described.

The coupling circuits are generally different for broadcast work because of the lower frequencies involved. One circuit commonly used with two-wire lines is shown in Fig. 201.

The procedure in setting up this circuit is first to determine the proper ratio of the circulating power to the power output of the antenna-coupler tank circuit. The basis for this ratio is not the same as for the plate tank circuit shown at the left of the diagram. The factors affecting this ratio will therefore now be considered in detail. The plate tank design was made to eliminate harmonics, and the antenna coupler will also reduce harmonics in proportion to the ratio of circulating watts (VA) to watts output of the tank circuit $\frac{VA}{\text{watts output}}$. The greater this ratio

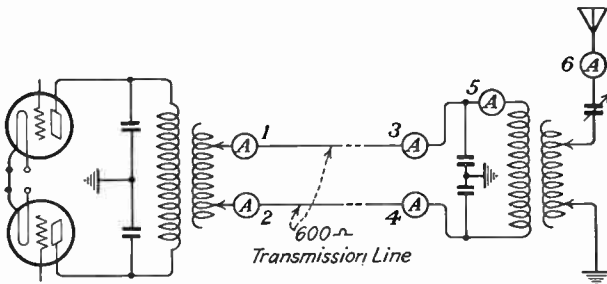


FIG. 201.—Coupling circuit commonly used on broadcast frequencies.

(especially with the center of the tank condensers grounded as shown in the figure), the lower will be the harmonic content. The limit to which this could be carried would be a point where the higher modulation frequencies are attenuated, owing to the large “flywheel” effect of the tank, and this point is far beyond the range where losses due to high circulating current become prohibitive.

Another consideration which affects this ratio is the impedance of the antenna. If the antenna is of high impedance, it makes a greater $\frac{VA}{\text{watts output}}$ ratio desirable, so that the step-up ratio of the coupling to the antenna coil does not have to be too great. In general a slightly higher C/L ratio is used than in the plate tank, but the voltage across the transmission line is comparatively low, so that the $\frac{VA}{\text{watts output}}$ ratio is lower. After this ratio is chosen, the value (C) of the tank condenser may be found from the equation:

$$C = \frac{VA}{W_o} \times \left(\frac{10^3}{2\pi f Z_o} \right) \tag{157}$$

where V = effective voltage across transmission line.

A = circulating current in tank circuit.

W_o = watts output.

Z_o = surge impedance.

The value of the inductance may then be calculated from the resonance equation, $2\pi fL = 1/2\pi fC$, substituting C as found above. It is best then actually to tune the tank circuit by means of an oscillator.

The following parameters of the coupling circuit are now known: antenna resistance; reflected resistance at the terminals of the tank which is equal to the surge impedance of the line, $R_o = 600$ ohms; capacity of the tank condenser. As the antenna will be tuned to series resonance, it can be assumed that a resistance equal to that of the antenna is connected directly across the antenna side of the coupler, as shown in Fig. 202.

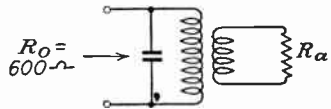


FIG. 202.—Equivalent circuit of antenna (R_a) across secondary of antenna coupler.

It can be shown that the mutual reactance between the two coils of the coupler is given by the equation

$$X_M = \sqrt{\frac{X_c^2 R_a}{R_o}} \tag{158}$$

where X_c = capacity reactance of tank condenser in ohms.

X_M = mutual reactance in ohms.

R_a = antenna resistance in ohms.

R_o = resistance equal to surge impedance of line in ohms.

The next step is to calculate the mutual inductance in microhenrys which will give this mutual reactance. This is accomplished by the equation for reactance:

$$X_M = 2\pi fL_M \tag{72}$$

As X_M and f are known, L_M can be found. L_M will be in henrys, so that multiplying by 10^6 will give microhenrys. The desired value of mutual inductance is then known, and it is simply necessary to know the physical dimensions and number of turns to use in the secondary. Empirical formulas 192 or 193, given in the U. S. Bureau of Standards *Bulletin* 74, are convenient for calculating this final stage in the design.

The procedure is then to tune the antenna for maximum antenna current by means of an oscillator using the series-antenna condenser, or a loading coil not inductively coupled to the antenna coupler. The tank condenser is then connected to the primary of the coupler, and the line is connected across the tank. The meters shown in the diagram are to check against reflections on the line, and to check the value of the mutual inductance (M) of the antenna coupler. If these line meters read the same on the input to the line as on the output, the impedance may be assumed to be matched. If the current readings are lower at the output of the line than at the coupler, it may be assumed that the antenna coupling is too great. If the readings are higher at the input, the coupling is not tight enough. If it is desired to operate the transmitter into a

dummy load, this load may be connected across the transmission line in the place of the coupler unit, but it must be non-inductive. A good type is the flat resistance-wire element woven on asbestos.

177. Application of the Antenna-coupling Unit.—As seen in Fig. 104(c) the output of the radio transmitter may be delivered to the radiating system through a transmission line and the antenna-coupling unit. Figure 203 is a schematic diagram of the entire antenna-coupling circuit. In the transmitter, at the left of the diagram, is the output circuit of the second power amplifier or last radio stage. The split condenser C_1 , C_2 tunes the primary of transformer T_1 . The secondary is tuned by C_3 and the coupling between the two windings is adjustable to permit the proper

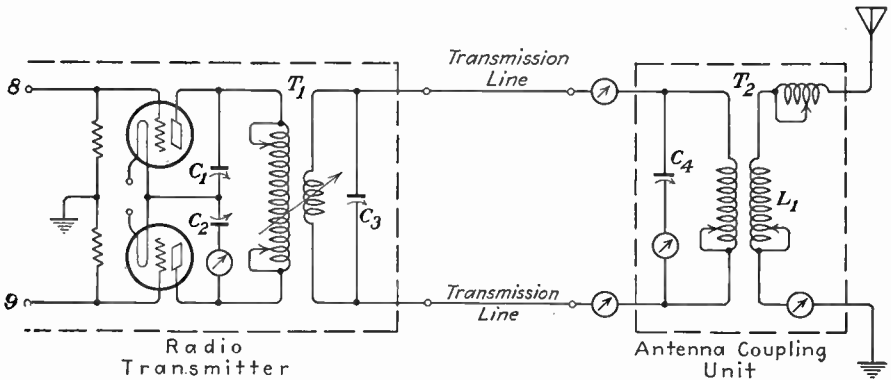


FIG. 203.—Schematic diagram of antenna-coupling circuit.

amount of resistance to be reflected into the primary circuit and thus to obtain the correct output impedance into which the tubes may work. T_1 may be considered an impedance-matching device for coupling a high-impedance circuit to a low-impedance one, because the combined plate resistance of the two tubes may be as high as 20,000 ohms, while the impedance of the transmission line may be 600 ohms.

The transmission line is merely a link circuit for connecting the transmitter to the antenna-coupling unit. In broadcast stations of 5 kw. or more, it is generally advisable to place the transmitter at some distance from the antenna. This serves a double purpose in that it gets the transmitter building and all associated structures away from the antenna so that there are no large obstructions near it with a resultant improvement of field patterns; and also the radio transmitter is removed from the intense electric field immediately surrounding the antenna, thereby reducing the possibility of radio-frequency feed back and instability in the transmitter.

The transmission line is terminated in the coupling transformer T_2 , which is again an impedance-matching device between the line and the antenna-ground system. The primary of the transformer T_2 is tuned

to the operating frequency by condenser C_4 . The secondary circuit is tuned by means of coil L_1 and the capacitive reactance of the antenna-ground system. The proper matching impedance for the line is obtained by adjusting the inductance of the secondary L_1 of T_2 , thus changing the mutual inductance and the amount of resistance reflected into the primary circuit. Both windings of T_2 are made of very few turns of large-diameter copper tubing wound on a large-diameter form. This is done to reduce the distributed capacity between turns and to have as little copper as possible, because as a rule the radio-frequency currents are quite high in both circuits. When properly adjusted, all the circuits are tuned to resonance, and the current in each wire of the transmission line is approximately the same. This condition indicates a proper matching of impedances and satisfactory operating results.

Since the output circuit of a tube transmitter must be connected into the proper load impedance to insure the maximum absorption and transfer of power by and to the radiating system, and also a minimum of wave reflection, both the transmission line and antenna impedance must be properly matched.

Where the output of a transmitter is to be fed into a transmission line, the line impedance must be carefully calculated to accommodate the impedance at both the input and the output of the line. This matched impedance of a line is sometimes referred to as a *surge* or *characteristic impedance*.

In the high-frequency transmission lines the characteristic impedance of the line must be kept as low as possible, approximately 500 ohms.

The Radiating System.—The antenna and the ground constitute the radiating system of the broadcast station, and this system coupled to the coupling unit and transmission line is the radio-frequency load into which the transmitter operates or delivers power. The antenna forms one plate of a condenser and the ground the other, the intervening space being the dielectric. This large condenser in series with the secondary of the coupling transformer and the loading coil, which tunes the whole circuit to series resonance, constitutes an oscillatory circuit with some value of internal resistance. This circuit which is tuned to series resonance at the operating frequency is set into oscillation by the voltages induced in it from the transmitter, and the resulting current flowing through the antenna resistance causes the radio-frequency power delivered by the transmitter to be dissipated in the form of heat and radiated energy.

Typical radiating systems in use are roughly divided into two general classes, namely: those used for marine transmission, and those used for broadcast transmission.

Those used for marine transmission on both land and sea are generally of the horizontal types known as the inverted L- or T-type antennas.

These antennas may use a number of wires in the horizontal section to increase the capacitance to ground, particularly in connection with the spark-type or damped-wave transmitters. The definite trend, however, for marine antennas is the single-wire T or inverted L type of radiator.

In broadcasting, the antenna systems have also been of the same type of horizontal T or inverted L type of radiators, but mostly of the single-wire variety. The definite trend, however, toward low-angle radiation in broadcast transmission is gradually eliminating the use of the old horizontal type of radiator in favor of the more efficient vertical type of radiator. These various types will be discussed in detail later.

178. Construction of Short-wave Antennas.—The different commercial communication companies have different specifications for the details of transmission lines and antenna construction, but the following are good workable values. Transmission-line poles should be at least 11 to 12 ft. above the ground. The cross arm should project from one side of the pole for a distance of about 4 ft., so that the wires of the line will not have the pole between them. The reason is that the pole is generally a poor dielectric, and the line should be kept away from it and the cross arms. Insulators which hang under the cross arm will be less apt to break than any fixed upright type. Insulators about 10 to 12 in. long of a high-grade porcelain or *Pyrex* glass will generally be satisfactory for transmission-line use at powers up to 20 kw. and over. Insulators with large metal-end caps should be avoided for transmission-line purposes, because the added lumped capacity will change the impedance of the line at this point and will cause reflections.

If the transmission-line poles are about 100 ft. apart, it will be found that No. 6 or No. 8 medium hand-drawn copper wire will stand the strain and is entirely satisfactory. The spacing for a 600-ohm line will be 12 in. and 9.6 in., respectively, for the above sizes of wires.

If the pole height suggested above is used, the line will be approximately 10 ft. above the ground, which is sufficient. The line impedance must be matched at the antenna. The method of accomplishing this for a doublet has been described, and for other types the "line-as-a-transformer" method, described above, is often convenient. This matching circuit may be placed vertically on one of the poles at a point where the antenna is to be fed.

A convenient method of matching the impedance of the line, and also of tuning the antennas to resonance, is by means of a small pick-up loop of wire with a thermoammeter arranged on a slider, so that it may be pulled along the transmission line, as shown in Fig. 204. There is no metallic connection between the line and the loop of wire. The coupling between the loop and the transmission line must be constant.

If the antenna is out of tune, it will not be a resistance load for the line, and there will be reflections or standing waves on the line. As the

loop slides along the line, there will be peaks of current and low readings one-quarter wave apart. First, the antenna is tuned until the least difference is found between the maximum reading and the minimum. If the antenna is a doublet, its length will have to be decreased one inch at a time, until the best length is found. If the antenna is changed to a directional antenna of the long-wire type, V or diamond, it will generally be possible to make this adjustment on one of the poles by means of a shorting bar.

After this adjustment is made, there will still be reflections on the line, because the degree of coupling will probably not be exactly right to match the line impedance. The next step is to vary the percentage of coupling. If the antenna is a half-wave doublet, the coupling will be

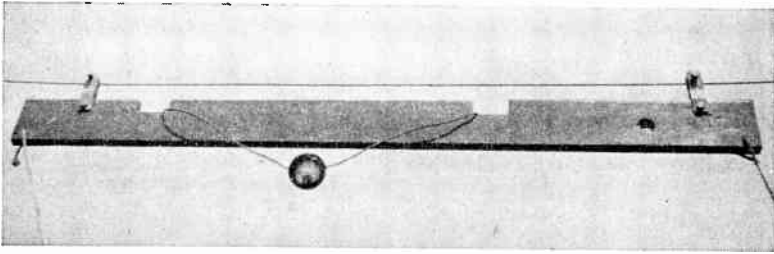


FIG. 204.—Scheme for reading current on doublet.

across approximately 20 per cent of the length, but this is only a starting point, and the exact distance to tap the feeder lines on each side of the center of the doublet should be determined by trial with the loop and a radio-frequency ammeter to indicate optimum matching and tuning. The size of meter to use in the pick-up loop depends, of course, on the power of the transmitter and the amount of coupling between the line and the pick-up loop. *In actual practice it is generally impossible to eliminate entirely all traces of standing waves*, but losses from radiation will not be great if the variation between maximum reading and minimum reading is not more than 5 per cent.

Another method of tuning the doublet is to place a radio-frequency ammeter at the current maximum, which is the center of the doublet, and read it with field glasses, lowering the doublet to change the length. This method would be preferable for the current-standing-wave and voltage-standing-wave feed systems described before, as the loop pick-up method, of course, could not be used on these lines. With these standing-wave feed systems it is only necessary to tune the series condensers until maximum current is secured in the line ammeters.

179. Ultra-high-frequency Antennas.—Because of the high frequencies involved, it is possible to use small antenna arrays, and to effect directional transmission by the use of wave directors and reflectors. Both Hertz and Marconi (grounded) radiators are successfully used, the type selected depending upon design contingencies.

The formula

$$\text{Length (feet)} = \frac{468,000}{\text{kilocycles}} \times K \quad (159)$$

for the half-wave antenna applies equally well for ultra-high-frequency antennas, where K is a constant depending on the loading effect of surrounding objects on the antenna. This factor can ordinarily be taken as 0.95 to 0.97.

The usual antenna erected for a fixed station operating in the band from 7 to 10 meters consists of a varnished wooden pole 25 to 40 ft. high which supports a 10- to 14-ft. voltage-fed half-wave radiator fed by a two-wire transmission line, the antenna being run up at an angle (nearly vertical) to the pole. The transmitter should be located within 150 ft. of the radiator and should be so placed that there will be no sharp turns in the transmission line.

A satisfactory antenna for 56 megacycles operation requires exact measurement of the half-wave radiator which may consist of a transmission line (two wires) 4 or 8 ft. long connected to two 8-ft. horizontal radiators extending out in diametrically opposite directions. An antenna ammeter may be inserted in one of the transmission lines to indicate resonance and the amount of current.

If possible, the antenna should be more than 30 ft. high above the ground; 50 ft. is even more satisfactory for an extended range.

180. Broadcast-transmitting Antennas.—In Sec. 162, 163 on propagation of broadcast frequencies under Sky and Ground Waves, it was shown that the local area within the range of the ground wave is becoming the important factor in calculating the service area of broadcast stations, and less attention is being given to Heaviside-layer reflections because of their instability. Therefore, broadcast-antenna design has been directed toward increasing the percentage of power radiated in the ground wave.

It is generally desired to have a non-directional field in the horizontal plane, and for this reason quarter-wave vertical antennas have been used for some time for broadcast purposes.

The quarter-wave vertical antenna will be used for comparison with newer antenna designs so that it is desirable to define several of the parameters of the quarter-wave radiator. The effective height of a quarter-wave vertical antenna is approximately 0.64 times the physical height. This effective height may be considered as the point at which all current flow can be considered to be concentrated and to secure the same result as the actual distributed condition. The *meter amperes* for a given frequency is this effective height in meters multiplied by the antenna current in amperes (I), and the received signal from the direct ray is directly proportional to this factor ($H_{eff}I$). This is not a

measure of power input to the antenna, which is, of course, determined by the product of total antenna resistance measured at the base and the square of the antenna current at the base. The radiation resistance is a function of the effective or useful part of an antenna system, and it is generally desirable to have it high, that is, as high a percentage of the total resistance as possible. The total radiated power is the radiation resistance multiplied by the square of the actual current at the base (I^2R). As pointed out above, a large part of this actual radiated power is wasted in the sky wave.

From this discussion it is evident that a good starting point is to increase the effective height thus increasing the meter amperes radiated. This is accomplished by making the vertical height greater than one-quarter wave. It was shown in an earlier section that this can be carried almost to the height of a half wave length and still tune to zero reactance by means of a series condenser.

It has been shown¹ that the optimum power efficiency for direct-ray reception will occur at a physical height of the transmitting antenna of 0.62 wave length. This is due to the fact that the next half-wave standing on the wire would be out of phase and would not add to the direct ray. The effect would be that described for long-wire radiators at short waves.

Ballantine shows that a great deal more of the radiated power is concentrated in low-angle radiation. As was shown in the development of a quarter-wave radiator from a half-wave doublet (Fig. 181), it may be assumed that the *ground* as a large capacity replaces the lower section of the wave or that the "image" of the upper section is in the earth. Then as the height of the antenna is increased above one-quarter wave, this image becomes of such space phase as to cancel the high-angle radiation without lessening the direct ray.

It is often found expensive to make towers as high as would be required for 0.62 wave length, so that some of the advantages of this design are used in a design slightly less than half a wave, down to about three-eighth wave length. It can easily be seen from curves of voltage distribution that the impedance into which the coupling unit must work will be lower as the antenna height is made either more or less than one-half wave length. If it is desired to reduce cost, it is possible to use some value between one-half wave length and three-eighth wave length. Under these conditions there will be an increase in ground wave as compared to a one-quarter wave-length antenna, and the impedance will still be low enough to make coupling easy.

Most of these designs are carried out in single steel towers pointed at the top and bottom and standing on a single heavy insulator. They are held vertical by means of guys broken up by insulators. When the tower itself is used as the antenna the structural members must be care-

¹ BALLANTINE, *Proc. I.R.E.*, December, 1924, p. 833.

fully bonded (connected together electrically) or special conductors must be run up one corner of the tower and bonded to the tower to reduce resistance losses.

If such an insulated tower is located near an airway, or if it is desired for other reasons to put lights on the tower, the power to operate the lights may be fed to the tower through a low-pass filter system which does not allow the radio frequency from the tower to back up into the power line. Such a filter circuit is shown in Fig. 205.

181. Ship Antennas.—The usual antenna system used with marine transmitters in the intermediate- and high-frequency bands is the Marconi or grounded radiator, in contrast to the Hertz or ungrounded radiator. The Marconi radiator is resonant to any odd multiple of a

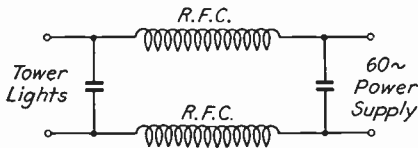


FIG. 205.—Low-pass filter for use in line to antenna tower lights.

quarter-wave length. This means that it operates normally as a quarter-wave antenna, or it may be operated on any odd harmonic of this wave, namely, as a three-fourth- or five-fourth-wave antenna. Owing to physical limitations, however, the

antenna used is generally a quarter-wave antenna. The quarter-wave transmitting antenna may be efficiently loaded up to about twice its resonant wave length by the insertion of antenna inductance, or it may be shortened down to about three-fourth of its resonant wave length by the use of a series condenser.

In the practical design of a Marconi radiator, allowance must be made for the coupling inductance necessary in the transmitter, which means that the resonant wave length of the wire from the remote end of the antenna to the antenna-terminal connection on the transmitter should be about 25 per cent lower than the lowest working wave length.

For example, to work on a minimum wave length of 560 meters, which is the low end of the marine band, a quarter-wave antenna could be 140 less 35 meters, or approximately 105 meters long. Converted to feet this equals 344.5 ft. Allowing 50 ft. for the lead in, this leaves an optimum length for the flat top of approximately 300 ft. The length of the lead in and flat top combined could be reduced to a minimum of approximately 150 ft. on small vessels, and good intermediate-frequency-operation would still be possible.

On large vessels the natural frequency of particularly long flat tops may be reduced by connecting the lead-in wire to the center of the flat top, thereby using the antenna as a T type.

The insulation of the ship antenna must be adequate, owing to the tendency of smoke and salt air to form a conducting surface over the insulator. The end or ends most remote from the transmitter are where the highest voltage strain exists and must be well insulated.

182. Airplane Antennas.—There are a number of different forms of antenna which are used on airplanes, the type used depending on the service to which it is to be put and the type of plane in which it is installed. Generally speaking airplane antennas divide into two classes, namely, the fixed or rigid type and the trailing-wire type. A number of different factors function to determine the selection of the antenna to be used on a given plane. It may be necessary to use more than one antenna under

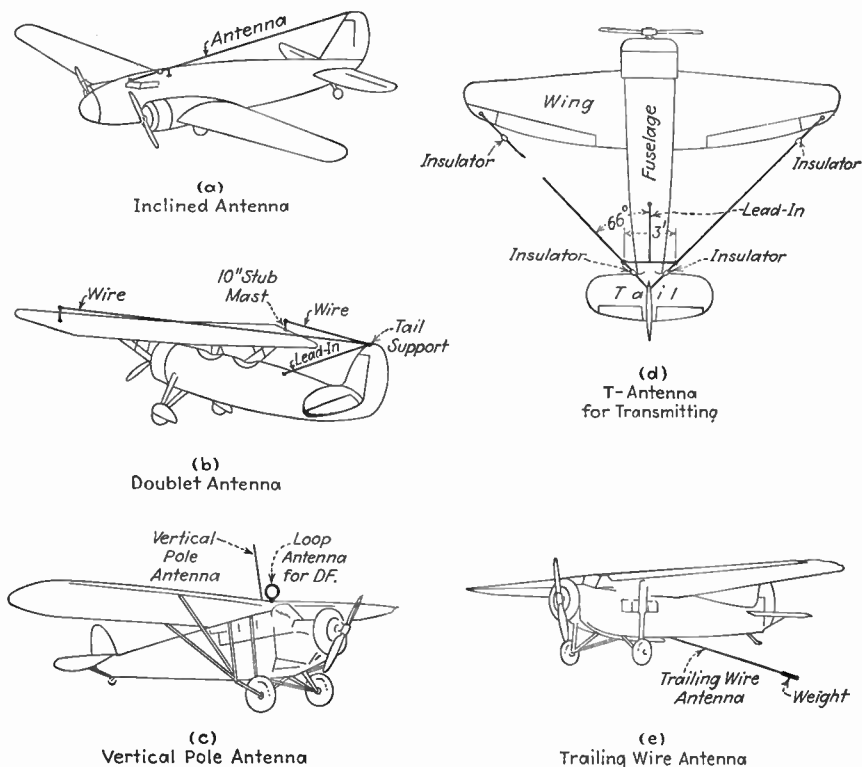


FIG. 206.—Types of airplane antennas.

special circumstances, a beacon receiver utilizing one of several types, a direction finder utilizing a loop and, if transmission from the plane is to be accomplished, still another set of provisions must be fulfilled. All of the antennas described, and referred to as being illustrated, are shown in Fig. 206.

FIXED TYPES.—*a.* The inclined antenna shown in (a) is supported between the tail surface and the forward part of the fuselage. It may or may not be supported on the forward end by a short stub mast 10 in. or 1 ft. high.

b. The doublet antenna may be erected on an airplane as shown in (b) using the ends of the wings and the tail for supports. If desired, a short

mast about 5 ft. high may be erected in the rear of the fuselage just forward of the tail to which the antenna may be fastened instead of to the tail surfaces. When this is done, the antenna lead in may be brought down through the hollow supporting mast.

c. The pole antenna consists of a hollow streamlined metal rod from 5 to 8 ft. in height which is conveniently mounted on the top of the fuselage as shown in (c). It must be thoroughly insulated from the supporting structure. Owing to its limited dimensions, it is not suitable for transmitting (except possibly ultra-high frequencies) and is mostly used for beacon reception.

d. The horizontal single-wire antenna parallel to the fuselage, or a *transverse antenna* parallel to the wings, may be used in certain installations for reception.

e. The T-type antenna largely used for transmitting is shown in (d). When it is desired to tune the transmitter over a wide band of frequencies, say from 3,000 to 6,000 kc./s., an antenna having a capacity of not over 200 mmf. should be used. The natural period of such an antenna may be from 85 to 95 meters. If frequencies up to 9,000 or 10,000 kc./s. are used, the capacity of the antenna should not exceed 175 mmf., the natural period then being approximately 50 meters.

TRAILING-WIRE TYPES.—The trailing-wire type of antenna is illustrated in (e) and is suitable for transmission as well as reception. It consists of a wire which may or may not be weighted at the end and which trails out behind the airplane while in flight. It must be reeled in when a landing is to be made. In addition to the reel a *fair-lead* is provided which guides the wire out of the fuselage, and also serves to make contact with the wire. If a weighted wire is used, the fair-lead is large enough in diameter so that the weight can be pulled through it into the fuselage when the antenna is reeled in. A cylindrical weight about the size of a large fountain pen is used to weigh down the trailing-wire antenna. Both the reel and the fair-lead must be insulated from the fuselage.

The length of the wire depends entirely upon the antenna constants desired and can be easily controlled.

When the trailing-wire antenna is supplied, it is sometimes the practice to supply a portable mast which may be set up for emergency purposes should it be necessary to operate the radio from the ground, the mast being used to support the end of the antenna remote from the plane. This end is commonly referred to as the *out end*.

On high-speed transport planes a gain in speed of as much as 7 m.p.h. has been secured by substituting an unweighted trailing-wire antenna for the mast-type antenna. The wire which for general purposes is 25 ft. long trails out straight behind without the weight being necessary.

183. Ground Connection and Counterpoise.—In considering the ground connection it must be remembered that the ground forms one

plate of a huge condenser of which the antenna forms the other plate when the Marconi quarter-wave radiator is used. On steel ships the ground connection is made to the hull, or, if it is a wooden vessel, a large copper plate several square yards in area is fastened to the hull below the water line and a connection made thereto.

At land stations, the ground connection is made by a radial system of wires buried in moist earth so that connection is made over a considerable area. In this way the dielectric loss caused by poor connection to the ground is kept low, and the service range of the station is measurably increased.

For receiving purposes, if connection to a transmitter ground is not possible, a connection to the water-pipe system of the building is in most instances satisfactory. Connection to the steam-heating and gas-pipe systems should be avoided where possible. Many receivers powered directly from the electric-lighting lines do not require a separate ground connection, as sufficient coupling to the ground is secured through the power lines and power-supply apparatus.

Where a good ground connection is not feasible or possible, a counterpoise is used.

Counterpoise.—One of the most important losses due to antenna resistance is caused by dielectric losses in poor ground connections. This is especially true where stations are located on dry, sandy, or rocky land, and is due to electron displacement in dielectrics traversed by the electric field set up by the antenna. It uses up useful energy in a useless action. In other words, if this loss is reduced, more energy is left for radiation purposes.

To reduce this loss a so-called *counterpoise* is used at some stations, which consists of a network of wires, erected a few feet above the ground but insulated from it and stretched out over the ground in a radial form with the radio-transmitter house as the center.

Theoretically the counterpoise forms one plate of the condenser of which the antenna is the other plate. The entire system of antenna, counterpoise, and ground may also be considered as two condensers in series, the antenna and moist ground far below the surface forming the outside plates while the counterpoise forms the center plate of the two condensers.

Where a counterpoise is erected, the same care should be taken to insulate it as is taken with the antenna. Supports should be as few as possible, and wooden stakes should be kept out of the field formed between the counterpoise and the antenna, as wood has a high dielectric power loss. All connections in the counterpoise should be soldered as in the antenna.

On aircraft and automobiles the metal chassis forms the counterpoise, and for this reason aircraft are carefully bonded to prevent sparking between poor joints in the counterpoise.

184. Receiving Antenna.—It may be that receiving antennas do not receive the attention deserved, but this is probably due to the fact that receivers have been improved to such an extent that elaborate receiving antennas are generally not necessary at broadcast frequencies. The only obstacle to good reception which can be overcome by a special antenna is interference due to pick-up from electric appliances, and the best cure for this seems to be to use a low-impedance, shielded-line lead in with a coupling transformer at each end. The cable may be a special lead-covered cable, or a wire covered by Belden Braid. The antenna coupler may be connected to one end of the antenna, the circuit being as shown in Fig. 207.

As stated before, the polarization of the propagated wave is vertical for long waves. From this it would be expected that a maximum signal would be received on a vertical antenna for a given length of wire.

Many different types of antennas have been used for receiving at short waves, ranging from a simple doublet to very elaborate directional

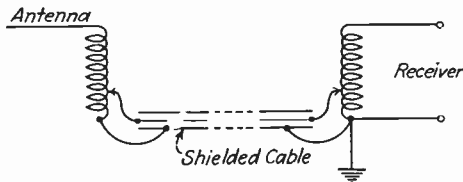


FIG. 207.—Receiving antenna coupling system.

designs. There is undoubtedly a very decided *gain* in the ratio of the received signal amplitude to background noise with directional receiving antennas, besides eliminating interference from signals from other directions. It can be seen that, from the receiving standpoint, gain must be considered in comparison with background noise, as in any receiver additional gain can easily be secured with more amplification, but the background noise is increased at the same time. For this reason a directional antenna has advantages over using additional amplification in the receiver.

In the discussion of polarization at short waves, it was mentioned that a horizontal antenna seems to have some advantage over a vertical arrangement. It is possible to arrange any of the directional-transmitting antennas as receiving antennas. However, if the RCA broadside arrangement is to be so used, there are generally considerable changes necessary in the design. The network is placed horizontally instead of vertically, and the short doublets act as collectors feeding a line. In series with each side of each doublet are placed small condensers. These are to assist in keeping the wave on the wire in phase with the wave in space. They reduce the loading effect of the short doublets. In this

way it is possible to increase the length of the antenna to several wave lengths.

The directivity pattern is with the line and not broadside as with the transmitting design. This receiving antenna is a development from the original Beverage long-wave antenna. The line and collector system is terminated at the end opposite the receiver with a non-inductive resistance equal to the characteristic impedance of the line, which runs from the antenna to the receiver.

Either the V or diamond types seem to have advantages as receiving antennas because of their simplicity. It should be mentioned that the same care should be used to match impedance at the line in a receiving antenna as with a transmitting antenna. Any of the types of transmission line described above may be used; however, it is customary to transpose the line about every 5 ft. to balance out pick-up on the line, and special porcelain insulators are made for this purpose. Where interference is particularly bad, the concentric-tube line might prove the best.

In the discussion on propagation at high frequencies, it was pointed out that certain frequencies are reflected best from the Heaviside layer at certain times of the day, but it will be found that, even with these optimum frequencies, there is still considerable fading at times. It has been noticed that fading does not occur at the same time on antennas spaced several wave lengths apart. Advantage is taken of this fact by using two or three separate antennas, each antenna going to a separate receiver. The outputs of the three separate receivers are then combined by special circuits depending on whether telephone or telegraph transmissions are being received.

Coil or loop antennas for receiving purposes have taken various forms, depending on the frequency and the physical space available. At the present time, loop antennas are particularly used in direction finding, both for ships at sea and for airplanes. A loop antenna is generally composed of a wooden vertical frame arranged to rotate on a pivot at the base and to hold the turns of wire in a vertical plane. The directivity is quite sharp and is at right angles to the axis of the coil or loop turns.

As used in direction finding, the loop is generally enclosed in a shield which surrounds the wire except that the shield does not form a closed ring, insulation being placed in a short section to prevent electromagnetic shielding as well as electrostatic shielding. The position of zero signal at right angles to the transmitting station is generally more sharply defined, so that it is used instead of the position of maximum signal. There are also several circuits used to make the direction-finding unidirectional. The use of the loop in direction-finder circuits is fully treated in Chap. XII.

Note: Antenna requirements for specific receivers are given in Chap. XI on Radio Receivers in connection with the description of the receiver as there given.

IV. ANTENNA MEASUREMENTS

185. Measurements on Broadcast Antennas.—*a. General Considerations.*—Although this section is labeled as applying to broadcast antennas, it applies, as a matter of fact, to all antennas of a similar design, such as ship antennas and antennas for general communication services.

Owing to the introduction of new antenna designs for broadcasting, certain factors, which must be measured, become much more important than formerly. Resistance measurements are generally made for two

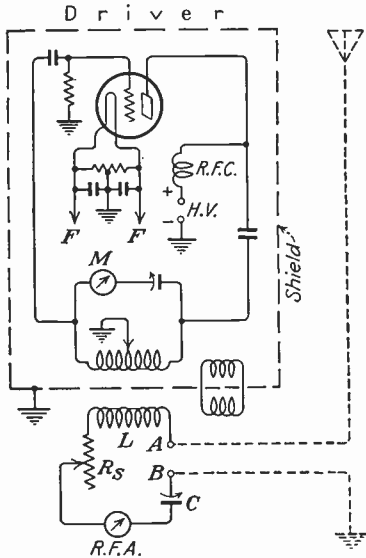


Fig. 208.—Circuit set-up for measuring antenna resistance.

purposes, first to be used in calculation of the antenna-coupling circuits, and second for the direct measurement of power output which direct method is required by the Federal licensing authority. This authority has made the rule that *the resistance which is to be used for this purpose is not to include the resistance of coupling circuits but simply the resistance from antenna post to ground.*

Although there are two methods of measuring antenna resistance, the resistance-substitution and the resistance-variation methods, the latter is chosen to be explained for the reason that better results can be expected by this method with the high-impedance antennas which are being developed and coming into common use.

b. A Simple Method for Measuring Antenna Post to Ground Resistance. The Federal licensing authority requires the use of the following apparatus in making these measurements.

1. A radio-frequency generator (oscillator) to cover the required range of frequencies. This generator to have a power output of 50 watts.
2. A wavemeter having an accuracy of 0.25 per cent.
3. A decade resistor box (non-inductive) having steps of units, tens and hundreds ohms resistance or the equivalent.
4. A radio-frequency galvanometer or milliammeter of an approved thermocouple type (0–100 ma.). This instrument must have an accuracy of 2 per cent.
5. An approved tuning condenser of approximately 0.001 mf. capacity.
6. A tuning inductance of 60 μ h.

Referring to Fig. 208, the measurement of the unknown antenna resistance R_a is made by connecting the antenna and ground across the terminals AB with the calibrated resistance R_s in zero position.

The resistance of the antenna may then be determined by the following procedure:

1. Set the driver into oscillation at the operating frequency, and couple it to the coil through the linking system as shown.

2. Carefully adjust the variable condenser C until a maximum deflection (resonance) is indicated on the radio-frequency ammeter RFA . The coupling between the linking circuit is to be sufficiently loosened so that the reading on RFA is about half scale, or 50 ma.

3. Start increasing the calibrated resistance R_s until the meter reads exactly one-half of its former value, or 25 ma. The amount of resistance introduced by R_s (in ohms) will then equal the antenna resistance R_a .

Note: This same instrument set-up may be used to measure the resistance of radio-frequency coils by connecting the unknown coil resistance R_x , instead of the antenna R_a across $A-B$.

Precautions: A limitation on the accuracy of the measurement is the existence of e.m.fs. electrostatically induced. In the deduction of the resistance it is assumed that the driver output power remains constant. The virtue of this method is that these stray e.m.fs. may be kept substantially constant during the measurement of the resistance of the circuit by proper grounding. They will invariably be altered by the substitution of the apparatus whose resistance is to be measured, but the resistance of the circuit is measured accurately both with and without the known resistance.

In order to keep these stray e.m.fs. unchanged, when R_s is in, and when it is out, of the circuit, particular care must be given to the grounding of the circuit. The shield of the condenser and the ammeter (particularly if it is a thermocouple with galvanometer) have considerable capacity to ground and are near ground potential. A ground wire, if used, must be connected either to the condenser shield or to one side of the ammeter. If it is connected to the high-potential side of the inductance coil, inaccurate results are obtained. The resistance R_s also must be inserted at a place of low potential, preferably between the shield side of the condenser and ammeter.

Furthermore, as previously stated, care must be taken that the coupling between the measuring circuit and the source is not too close. Otherwise the current in the source, and hence the e.m.f. E will vary somewhat when R_s is inserted. This will give incorrect resistance values which will depend upon the magnitude of R_s . Whether such an effect is present can be judged by opening and closing the measuring circuit and noting whether this produces a considerable change in the ammeter reading in the source circuit; or by repeating the measurement with reduced coupling. In order that the measurement can be made using very loose coupling, it is necessary either to have a source of considerable power, or to use a sensitive current-measuring device such as a thermocouple galvanometer.

c. Conventional Method for Measuring Total Antenna-circuit Resistance.¹

As regularly carried out at the U.S. Bureau of Standards by the resistance-variation method, a vacuum-tube oscillator is used as a source of undamped-wave e.m.f., and the current is measured with a low-resistance thermocouple galvanometer in series with the circuit being measured. The currents corresponding to given deflections of the thermocouple

¹ Essentially as given in *U. S. Bureau of Standards Circular 74*.

galvanometer are obtained from a calibration curve; or from the square law $d \propto I^2$ where d is the deflection if the meter follows this law close enough. Under these conditions the equation becomes

$$R = \frac{R_1}{\sqrt{\frac{d}{d_1} - 1}} \quad (160)$$

Where R denotes the resistance of the circuit without the added resistance, and R_1 is the added resistance introduced in the circuit.

Referring to the circuit diagrams in Figs. 208 and 209, several values of resistance R_1 , R_2 and R_3 are usually inserted in the circuit at R_s and the corresponding deflections obtained; the resulting values of R being averaged.

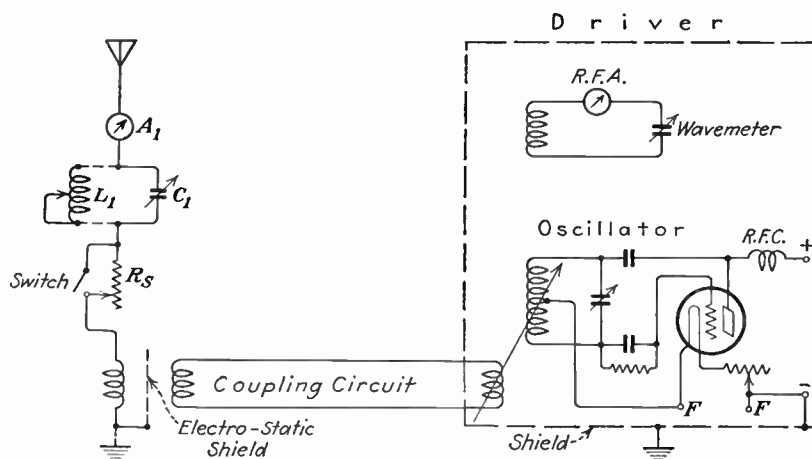


FIG. 209.—Conventional method for measuring total antenna-circuit resistance.

When the thermocouple galvanometer follows the square law accurately, the quarter-deflection method may be used, which eliminates all calculation. When the deflection d_1 is $\frac{1}{4}d$, the equation becomes

$$R = R_s$$

This method requires a variable-resistance standard such that R_s can be varied continuously in order to make d_1 just equal to $\frac{1}{4}d$. Practically the same method is used, if the resistance is varied by small steps, as in a resistance box, interpolating between two settings of R_s . Suppose that the three values of R_s are R_1 , R_2 , and R_3 , and the corresponding deflection d_2 is slightly less than $\frac{1}{4}d$, and d_3 slightly greater than $\frac{1}{4}d$, then

$$R_s = R_1 + \frac{d_3 - \frac{1}{4}d}{d_3 - d_2} \times (R_2 - R_3)$$

A practical measurement of the radiation resistance by the resistance-variation method using the circuit illustrated in Fig. 209, may then be readily made, using the equation

$$R_a = \frac{R_s}{\frac{I}{I_1} - 1}$$

where R_a = resistance of the antenna.

R_s = value of the added resistance.

I = current reading without the added resistance.

I_1 = current reading with the added resistance.

Procedure: 1. Adjust the antenna circuit to resonance with the normal operating conditions of the transmitter. The radiation ammeter (A_1) may be left in the circuit.
 2. Take out all R_s resistance by closing the short-circuiting switch.
 3. Start driver, and tune it to resonance with antenna circuit.
 4. Record current reading on radio-frequency milliammeter (A_1). This becomes the I value in the formula.
 5. Do not vary power in driver. Open switch and insert a small value of R_s .
 6. Record current reading on radio-frequency milliammeter A_1 . This becomes the I_1 value in the formula.

The above procedure may be repeated *with four or five different values of R_s* and the results averaged. Accurate results are more likely to be obtained in this way.

The power in the driver must be constant throughout the entire measurement, or the results will be inaccurate.

If a current-squared galvanometer or wattmeter is used in place of the radio-frequency milliammeter, the square roots of the readings must be used for current readings.

The resistance of the radio-frequency milliammeter A_1 , and any additional L or C in the antenna circuit, must be subtracted from the value of R_a , computed as described, in order to get the true R_a of the antenna system as it is under operating conditions.

d. Some Practical Consideration In Making Antenna Measurements.—The wave-meter should be coupled to the oscillator, not to the antenna-coupling circuit. The variable condenser C_1 must be of low-loss construction and must be calibrated. If the antenna is less than one-quarter wave length or over one-half wave length high, as referred to the operating frequency, a variable inductance L_1 will have to be substituted for variable condenser C_1 . This is shown dotted in the diagram. This condenser or inductance must be entirely away from all grounded shields or screens, and all regular coupling equipment must be grounded or removed. The coupler shown between the oscillator and the antenna may have two or three turns in the antenna side and the same in the link circuit. The diameter of these coils may be about $1\frac{1}{2}$ in., and the electrostatic screen may be made of strips of thin copper running lengthwise on the coil, with all strips soldered to a wire at one end of the coil. However, this wire should not make a closed ring. The decade resistance box R_s must be of a non-inductive variable type, such as are used with bridge circuits. The meter

to be used at A , may be a 100-ma. thermocouple meter, or a Weston thermogalvanometer which reads 115 ma. full scale. The latter meter reads current square.

Great care should be used to get the antenna exactly in resonance before reading. It is generally best to take readings at about 10 points, making a curve several kilocycles each side of the operating frequency. If a reading is desired at the fundamental, the condenser or coil is taken out of its circuit and the oscillator is then tuned to the antenna frequency, and a resistance measurement made using the procedure before described.

The errors due to the capacity of the inductance coil to ground are very serious for a high-impedance antenna, which is approximately one-half wave length high, because of the comparatively high voltages at the antenna terminal.

From the curve of Fig. 183, it can be seen what reactance to expect for any antenna if its fundamental is known. If this curve showed inductive-reactance values for the antenna, a variable condenser would be used in the above measuring circuit to tune it. Then, from the capacity value read from the calibrated condenser in the above measurement, a more exact calculation can be made of the inductive reactance of the antenna. For resonance we know

$$X_L = X_C \quad \text{or} \quad 2\pi fL = \frac{1}{2\pi fc}$$

The value of f and c are known from the data taken above, so that X_L , the inductive reactance of the antenna in ohms, may be calculated. The antenna impedance equation is then $Z_a = r_a + jX_C$, where r_a is the antenna resistance, and jX_C is the antenna reactance.

It may here be assumed that the resistance of the low-loss air condenser is zero, but, where the antenna has capacitive reactance, and it is necessary to use a loading inductance, the resistance of the inductance coil cannot be assumed to be zero. The resistance value should be known or measured and may be as high as 10 or 15 ohms. The same equation can be used if, to tune, inductance is added, but the antenna equation will then come out

$$Z_a = R_a - jX_C$$

186. Calculations on Antennas.—For a quarter-wave flat-top antenna, it is possible to calculate the fundamental wave length, λ_f , by the following empirical formula:

$$\lambda_f = 4.2[(\text{vertical height in meters}) + (\text{length of flat top in meters})] \quad (161)$$

The equation is for the L type, but for a T type the flat top on one side of the center is used. The factor 4.2 may increase to values as high as 6 for flat tops of large area.

Radiation resistance of a quarter-wave antenna may be shown to be

$$R_a = 1,580 \times \frac{h_e^2}{\lambda^2} \quad (162)$$

From this the total resistance of the antenna may be estimated. This is simply the radiation resistance (R_a), plus resistance losses in the ground, wire dielectrics, and any other small losses. In general, for one-quarter-wave broadcast antennas this will probably be from 5 to 15 ohms; for three-eighths, 35 to 75 ohms; for half-wave, 200 to 600 ohms; for five-eighths, 100 to 150 ohms.

In the design of broadcast antenna systems there are many important factors which must be considered which extend far beyond the scope of this text.

The most important of these, however, may be summarized as follows.

1. The radiating system must be designed to transmit low-angle wave radiation to reduce energy waste due to high-angle or sky-wave radiation. The concentration of the energy into a low-angle or ground-wave propagation increases the local field-strength pattern and reduces sky-wave radiation and subsequent heterodyne interference at remote points. Low-angle radiation is most effectively obtained by the use of a half-wave vertical antenna system. If the half-wave design is impractical for the frequency required, the three-eighth and five-eighth vertical systems may be used. The choice of either of the three systems is primarily dependent upon the frequency at which the antenna system is to be operated.

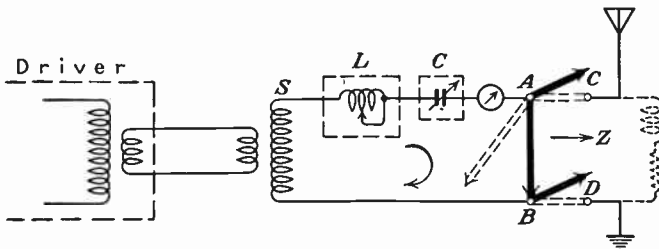


FIG. 210.—Suggestion for determining antenna impedance.

2. The antenna must be designed within the limits of definite impedance values so that the maximum degree of energy is absorbed and radiated by the system. This is an important consideration since high-impedance antennas introduce too many variables to enable stable operation. The impedance looking into the antenna system must therefore be very carefully determined before transmission lines and coupling units are designed. The antenna resistance does not usually exceed 300 ohms at any of the operating frequencies in the broadcast spectrum. Although it is possible to operate into antennas of high impedance with fair efficiency if the exact impedance can be determined, it is extremely difficult to design accurate coupling devices owing to the uncertainty of the exact antenna-impedance value.

Since an antenna system is a power-absorbing device in which the power is converted into electromagnetic-radiation energy, it should be apparent that maximum efficiency can only be obtained when the antenna impedance is carefully matched to the exciting impedance in accordance with the fundamental rules of impedance matching.

Designing engineers give this the most serious consideration before designing transmission lines and coupling units to operate in conjunction with the radiating system. By first calculating the reactance of the

unloaded antenna, it will also be possible to determine if the antenna is either highly inductive ($+j$) or highly capacitive ($-j$), which, therefore, makes it a relatively simple problem to determine whether the antenna should be loaded inductively or capacitively. It also provides the necessary information for the design of the transmission line and the coupling units.

An arrangement for determining the reactance looking into an antenna system is shown in Fig. 210. The inductance L is a low-loss variable coil, and the capacity C is a variable condenser with a calibrated capacity chart. The capacity in microfarads for each setting may be determined by referring to the capacity chart, and the reactance then computed for the particular setting by the ratio, $X_c = 1,000,000/2\pi fC$ ohms.

The procedure for determining the antenna reactance is then as follows: Close the links AC and BD , and open the link AB . This connects the antenna reactances in series with the circuit components, S , L , and C . Now with the driver oscillator adjusted to resonance at the operating frequency, and tuned to the point of maximum dip on its plate circuit direct-current milliammeter (unity power-factor adjustment), the calibrated variable condenser C is adjusted until a maximum deflection is obtained on the radio-frequency ammeter. This reading will then indicate that the entire circuit reactance has been reduced to zero. The condenser setting should then be noted, its capacity determined by referring to the calibration chart, and the capacitive reactance computed from the reactance formula. It is sometimes necessary to readjust the driver circuit after the antenna circuit has been tuned owing to circuit reflection effects. If, however, the linking circuit between the driver and the coil S is composed of only two or three turns, a readjustment of the driver circuit to unity power-factor conditions will not be necessary. The links AC and BD may now be opened, and the link AB closed. The driver is again set into operation, and the circuit LC tuned to series resonance by adjusting the calibrated condenser C . When a maximum deflection is noted on the radio-frequency ammeter, the condenser setting is again carefully noted, and the capacity chart referred to. The reactance at the new setting is then calculated as before. The difference between the two reactances will then determine the effective antenna reactance. Now, if it was found that the condenser C had to be increased to effect resonance, the antenna system may be looked upon as a $-j$, a capacitively reactive quantity, or vice versa. If standard calibration curves are provided for L and C , it would be a simple matter to determine whether the antenna reactance is a $-j$ or $+j$ value. Thus by taking the previously measured antenna resistance (Sec. 185) and the present reactance value $-jX$, the antenna impedance may be readily determined by the expression $Z_a = R_a \pm jX$.

CHAPTER VII

STUDIO ACOUSTICS AND APPARATUS

Any study of broadcasting must consider the subject of acoustics and sound control as these are the first factors affecting the broadcast program.

I. STUDIO ACOUSTICS

187. Distortion.—The never ending problem of the radio studio is distortion. The importance of this matter is evident because no amplifying apparatus, even though it may be almost perfect in design, can give faithful reproduction if the sound impressed upon its input is distorted. And this may take place in many ways. There are two ways in which the microphone itself may be the cause of distortion. One of these is by distortion of the wave front, consisting of a series of condensations and rarefactions in the air, or more simply, variations in air pressure, due to the shape and bulkiness of the microphone itself.

Such distortion is similar to the change caused in the form of a water wave when it hits a large immovable object such as a pier. If the immovable object is comparatively small and offers little resistance, such as the edge of a knife blade, the true form of the wave at that point in space is very little altered. Similarly, if it is desired to keep a microphone from altering the audio-frequency wave form striking its active surface, every effort should be made to approximate as nearly as possible the same relation between it and the sound wave as was present between the knife blade and the water wave. This condition becomes more difficult to fulfill as the audio frequency is increased and the wave becomes correspondingly smaller.

The particular type of distortion caused by the microphone cannot ordinarily be corrected when the microphone is in use. It is necessary, therefore, so far as possible, to take care of this in the original design of the microphone. Controlling factors are the size and shape of the microphone, which should be kept as small as possible and be streamlined to sound. In other words, an effort should be made to achieve acoustical transparency.

Another cause of distortion at the microphone is acoustical resonance. This results from the cavity formed by the diaphragm and the walls of its retaining ring. This effect and its remedy are discussed in detail in **Sec. 195** on the Condenser Microphone.

188. Reverberation. 1. *The Terms "Live" and "Dead."*—These terms enter frequently into any discussion of acoustics. If it is said that a room is dead, it is meant that there is little or no reflection of a sound wave from the walls of that room. This term may refer to only a particular section of a room, that is, a room may be only partly dead. An example of a truly dead position would be a location in interstellar space where any sound originating from a point would never return to it in the form of a reflected wave.

On the other hand, in a live room, a great deal of reflection takes place, and the sound may return to its source many times before it finally dies away. An example of a live room is one having highly reflecting tile walls. This term also may refer to only a part of a given room. Thus there are gradations between a very live room and a very dead room, depending on the *reverberation* characteristics of the room, that is, upon its ability to reflect sound waves.

2. *Reverberation time* is an expression of unit reverberation and is defined as follows: Assume a sound source to produce a sound intensity (energy) at a given point in a room of one million times the energy which would be just audible; if this source be stopped, the reverberation time is the time taken for the sound energy at that point to fall to a just audible value.

A formula, known as *Sabine's formula*, which is sufficiently accurate for most rooms, is expressed.

$$t = \frac{0.05V}{as} \quad (163)$$

where t = reverberation time in seconds.

V = volume of room in cubic feet.

a = average coefficient of absorption of these surfaces.

s = total area of absorbing surfaces in square feet.

It is thus seen what factors enter into making a room live or dead.

A microphone placed in an almost dead studio, that is, one having walls capable of a high degree of sound absorption (short reverberation time) will pick up the required sound only once, that being directly from its source. Contrary to what might be expected, the effect of this is usually displeasing. We are accustomed to a certain amount of reverberation, and, if this amount is not present, the sound affects us as being flat, shallow, and uninteresting. For the same reason it is very irksome to the artist to work under these conditions.

3. *The Absorption Coefficient.*—There is an ideal time lag (reverberation time) and volume proportion between the original and reflected sound striking the microphone. If this is achieved the resultant sound will be natural in quality, neither too dead nor too live.

The quantity relation between the original and reflected sound depends, among other things, upon the efficiency of the reflecting surface. Different materials possess reflecting properties in varying degrees, dependent upon their structure and surface. In a treatise on acoustics it is more usual to speak of the absorption coefficient of a material than

its reflecting efficiency. The *absorption coefficients* for a variety of materials have been measured at the U. S. Bureau of Standards, and the results, at frequencies from 128 to 4,096 cycles, are available upon request.

The value of the absorption coefficient becomes higher as the frequency of the sound wave increases. The ideal number of fixed absorption units present in a studio depends upon a number of things, the most important being the cubic content of the room, the number of performers present, and the instruments involved.

The time lag between the original and reflected sounds at the microphone is not important if the sounds are not comparable in intensity. If they are comparable it is best to keep the lag short. This may sometimes be accomplished by adjustment of the distance between the microphone and the reflecting surface. In practice, if the conditions surrounding the sound source and microphone are too live and the time lag is very short, the effect produced makes the reproduction sound mechanical and tinny. If, under the same conditions, the time lag is rather long, hollow and cavern-like effects are secured.

189. Microphone Placement.—In practice, the sound relation is very important, and it is usually the easiest to correct. In situations where acoustic conditions already existing must be met, such as in certain broadcasts remote from the studio, satisfactory reproduction can usually be secured if the following operations are carried out. The microphone should be placed as far as possible from any highly reflecting surface. As sound radiates from a given point, its intensity decreases as the square of the distance from the source. Thus, by increasing the distance between the sound source and the reflector, we achieve a decrease in the value of the reflected sound returned to the microphone. Another most important factor in the fight for good reproduction is the possibility of adjustment of the distance between the microphone and sound source. The volume relation between original and reflected sound, and hence the resultant characteristic of the microphone output, may most easily be influenced by experimentation with this distance. Decreasing the distance increases the proportion of original sound, and vice versa. Many times proper microphone placement affords the only solution to the problem offered by poor acoustics.

It has been stated that the absorption coefficient differs with the material under consideration; also that it increases with the frequency of the sound wave. Thus, if a studio is treated by applying sound-absorbing material to its walls, so that it is ideal for the rendition of a piccolo solo, which consists mainly of high audio frequencies, it may prove hopelessly inadequate for the proper reproduction of a duet between tympani and double bass. Also we may find that the studio which was ideal for a piccolo solo is unsuited to a band of 50 piccolos, as the added absorption

of 49 more people, and their effect in breaking up the reflected sound waves may deaden the studio to too great an extent.

Adjusting Studio Acoustics.—In adjusting the acoustics of a broadcasting studio, an absorbing material should be chosen whose action is efficient at all audio frequencies. Unfortunately this is something much easier to say than to do. Rock wool and monk's cloth are frequently used, as well as materials specially designed and prepared for this purpose. The amount of deadening material used should be readily adjustable to meet varying conditions. This may be accomplished by means of sliding or folding panels. Sometimes large screens of sound-absorbing or sound-reflecting materials mounted on rollers so that they may be moved to the most advantageous position have been used successfully.

Even with this help, compromises must be made in arranging (known as *setting up*) the artists before the microphone. By way of example, the absorption coefficient of materials is very low at the frequencies set up by a double bass. Owing to this fact a greater degree of reflection will take place for this instrument than for others of higher pitch. To prevent improper relation between original and reflected sound, the bass is placed closer to the microphone than other instruments of a higher frequency output. This makes the sound energy received at the microphone from this instrument proportionately large and, in order to secure musical balance, makes necessary the reinforcement of the high end of the audio spectrum by the addition of higher pitched instruments. Thus the normal concert proportions and arrangement of the instrumental sections of an orchestra are not necessarily normal to broadcasting.

The adjusting of studio acoustics and the setting up of the artists for broadcasting have become a highly specialized field which is taken care of by experts who are known as "production men."

Many experts believe it best to have a live and a dead end in each studio. The microphone is placed at some distance from the performers in the dead end of the studio and is encompassed on three sides by sound-absorbing material. The artists are placed in the live end. This simulates the conditions found in the usual concert hall, where the platform is live and the ear of the individual in the audience is surrounded by other persons who are acting as sound-absorbing surfaces. It is questionable, however, if the best place to hear music is in a concert hall.

Let us suppose that by means of careful design a "perfect" studio has been achieved. It is desired to broadcast a playlet having three characters, and for the sake of convenience each player is given an individual microphone. These microphones are placed fairly close together and their outputs are fed into an amplifying system. It may be found that this studio, which gave perfect reproduction when used with a single microphone, has now developed a cavernous echo. This is not due to the studio acoustics but to the fact that the sound intended only for micro-

phone 1 is also reaching microphones 2 and 3 at slightly different time intervals and volumes. The effect on the resultant output is similar to

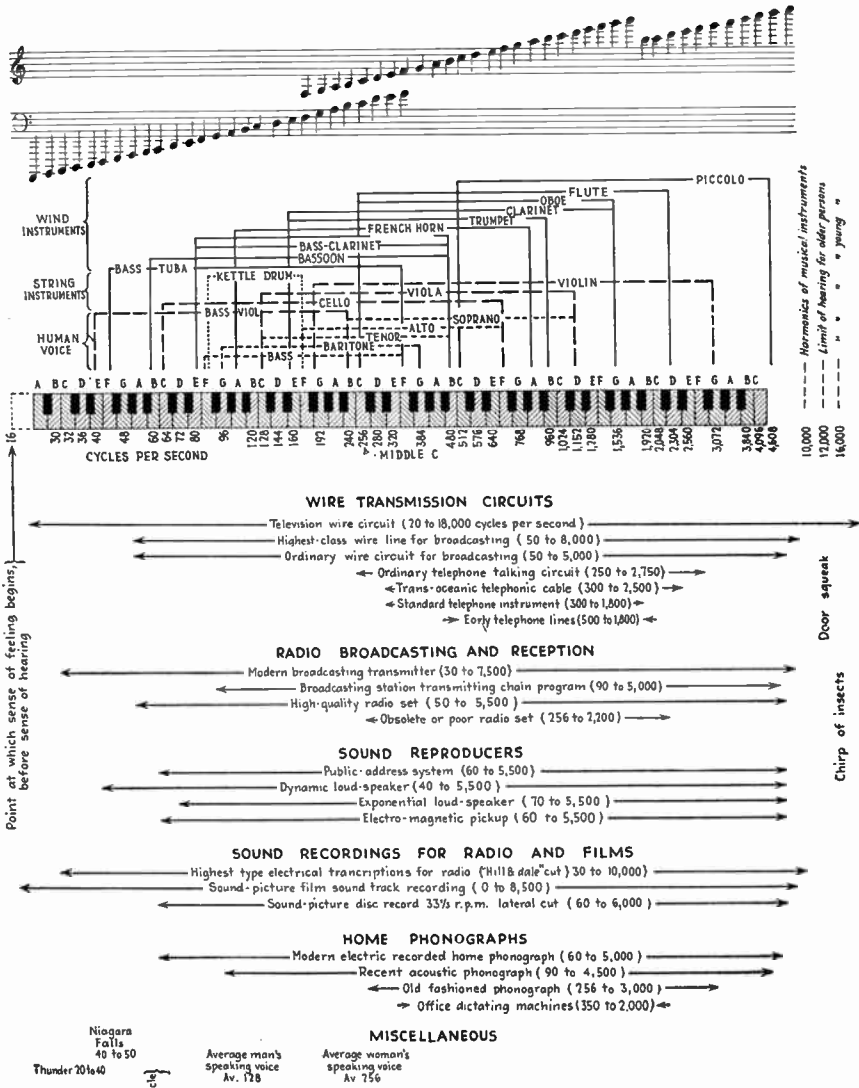


FIG. 211.—The audio spectrum and its relation to broadcasting. (Courtesy of Electronics.)

that which we should have if using one microphone in highly reflecting surroundings.

Because of this multi-pick-up effect, it is an excellent working rule always to use as few microphones as possible.

The acoustical properties displayed by different types of microphones will be taken up in conjunction with their electrical properties at a later point in this chapter.

190. Sound is divided roughly into two parts, namely, tone sensation and noise sensation. Tone may be graphically represented by a simple sinusoidal curve, while noise is irregular in character. A tuning fork gives the best example of a simple tone. Ordinary music is very complicated in character and consists of a simultaneous sounding of many simple tones together with complex noise waves. For example, the music emitted by a single violin string may consist of as many as six simple tones plus the noise of the bow scraping the string.

The audible spectrum describing the limits of human hearing is shown in Fig. 211. This graphical analysis of sound is well worth careful study.

191. Use of Logarithms.—By the use of the following information the student not instructed in the use of logarithms may, with some little coaching, be able to handle the decibel computations which appear in the following section.

The common logarithm (abbr. log) of any number is that power to which the number 10 must be raised in order to equal the given number. Hence it is necessary to raise 10 to the second power, or “square” it (10^2), to get 100; therefore, the log of 100 is 2. The log of 1,000 is 3, as we must cube 10 (10^3) to get 1,000. From this it follows that the log of any number between 100 and 1,000 must be 2 plus a decimal. For example, the log of 280. is 2.447158. The 2 is called the *characteristic* and the decimal .447158, the *mantissa*.

The characteristic of any number is *one less* than the number of digits to the left of the decimal point. If the number is less than 1, the characteristic is *negative* and *one greater* than the number of zeros between the decimal point and the first significant number. However, as negative characteristics can be avoided by a method to be suggested later, they need not concern the student for the present. The following tabulation shows the relation between any given number and its characteristic in accordance with the rule just mentioned for positive characteristics:

Number.....	1500	150	15	1.5
Characteristic.....	3.	2.	1.	0.

It must be borne in mind that the characteristic is not the whole logarithm but only the first part of it. The second part, to the right of the decimal point, which is the mantissa of the log, must yet be found. It is for this purpose that a book of logarithmic tables is needed. A logarithmic table is a table of mantissas, as the characteristic is found without a table. Thus the mantissa of a number is found in the table of logarithms, in accordance with directions accompanying the book of tables and which are not given here because they can only be followed intelligently with a book of tables available. The following rule must always be remembered in dealing with logarithms: *The mantissa is always a positive number.*

A very incomplete tabulation of mantissas is given below so that the student may make a rough check on the pad formulas to be taken up later.

Significant Digit	Mantissa
2.....	3
3.....	.477
4.....	.60
5.....	.7
6.....	.78
7.....	.845
8.....	.903
9.....	.954
12.....	.08
14.....	.146
15.....	.176
18.....	.255

The *antilog of any logarithm* is simply the number from which the logarithm was derived. It can be found by working backward, or by a table of antilogs. The mantissa gives the number, and the characteristic places the decimal point. For instance, we wish to find the antilog of the logarithm 3.176. By looking up the mantissa (.176) we find that this comes opposite the number 15 in the above table. This gives us the significant figures of the number, but we must place the decimal point by inspection. As the characteristic is 3, we know that the number must have four integers to the left of the decimal point. Hence it is 1500. The proof is that the log of 1500 is equal to 3.176.

A little practice will give familiarity and assurance. A complete log table can usually be found in the appendix of any textbook on trigonometry, although a book devoted exclusively to mathematical tables is better.

192. The Decibel.—Inasmuch as the ear responds logarithmically to variations in sound intensity, any unit used to compare sound levels must, in order to be practical, vary logarithmically. Such a unit is the *decibel* which is one-tenth (deci) of the international transmission unit known as the *bel*, named in honor of Dr. Alexander Graham Bell, the inventor of the telephone. The abbreviation for decibel is db.

The decibel expresses the logarithmic ratios of powers, voltages, or currents in a transmission system. As such it is very useful in expressing gains in amplifying systems and losses in attenuation networks, and in comparing the electrical output of any piece of equipment with a standard.

A listener with acute hearing powers would become aware of a change in sound intensity of a single sustained note when the level is varied 1 db, but not before. The average person would not notice a change in a radio-program level of less than 3 db.

The decibel is expressed mathematically by the equation,

$$\text{db} = 10 \log_{10} \frac{P_2(\text{larger power})}{P_1(\text{smaller power})} \quad (164)$$

By always placing the larger power value on the top of the ratio as the numerator, the result will always be greater than one, and the characteristic of the log of the ratio will always be zero or a greater positive value.

Thus the use of complicated negative characteristics is avoided. It is obvious in each problem whether a gain (+) or a loss (-) in decibels results. The negative sign can be inserted before the decibel value if a loss is indicated. This will become apparent as the problems are worked out.

A table showing the relation between power ratio and the decibel, and voltage or current ratio and the decibel, is given below. If the proper ratio is known, it is possible, by referring to this table, to ascertain the exact or approximate result in decibels. Likewise, if the decibel value is known the exact or approximate required power, current, or voltage ratio may be found.

TABLE IV.—RELATIONSHIP OF RATIOS TO DECIBELS

Power ratio	Transmission units in decibels (db)
1 ($= 10^0$)	0 ($= 10 \log_{10} 1$)
1.259 ($= 10^{0.1}$)	1 ($= 10 \log_{10} 1.259$)
10 ($= 10^1$)	10 ($= 10 \log_{10} 10$)
100 ($= 10^2$)	20 ($= 10 \log_{10} 100$)
1,000 ($= 10^3$)	30 ($= 10 \log_{10} 1,000$)

Voltage or current ratio	Transmission units in decibels (db)
0.001	-60.00
0.005	-46.02
0.01	-40.00
0.05	-26.02
0.1	-20.00
0.2	-13.98
0.5	-6.02
1.0	0.00
1.5	3.52
2	6.02
5	13.98
10	20.00
20	26.02
50	33.98
100	40.00
500	53.98
1,000	60.00

Note: In the practical notations which follow, \log_{10} is simply written as *log*, as only logs to the base of 10 are used in this text.

Example 1: It is desired to increase the distortionless volume of a radio receiver. The output tube in use is a -45 type supplying a maximum power to the loud-speaker of 1,600 mw. Consideration is being given to changing this -45 tube to a type -50 tube which has an output of 4,600 mw., almost a 2.9 times increase in power. Anyone inexperienced in the use of the decibel would most likely consider an almost threefold increase in sound level to result from such a change of tubes. This however is not true, as can easily be shown:

$$\text{db} = 10 \log \frac{P_2}{P_1} = 10 \log \frac{46}{16} = 10 \log 2.87$$

$$\log 2.87 = 0.46; \quad 10 \times 0.46 = 4.6$$

Therefore,

$$\text{Gain} = 4.6 \text{ db}$$

Such a change would only be noticeable by direct comparison with the previous power level and would not be sufficiently great enough to warrant the expense involved.

Example 2: If the type -50 tube in the preceding example is not satisfactory, what power increase would be necessary to effect a noticeable increase in sound level? A change of power to give a gain of 15 db might be tried. Let X represent the new power. The change is to be made from the 1,600 mw. supplied by the -45 tube.

$$\text{db} = 10 \log \frac{P_2}{P_1} = 10 \log \frac{X}{1,600}$$

Therefore,

$$15 = 10 \log \frac{X}{1,600} \therefore 1.5 = \log \frac{X}{1,600}, \quad \text{antilog } 1.5 = 31.6$$

Therefore,

$$\frac{X}{1,600} = 31.6 \quad \text{and} \quad X = 1,600 \times 31.6 = 50,560 \text{ mw.}$$

Ans.

$$X = 50.5 \text{ watts}$$

Such a large increase in power would necessitate the use of a powerful push-pull amplifier which would be rather costly.

Example 3: The necessity of handling large computations in multiplication and division is avoided when the gain in an amplifying system is figured directly in decibels.

It is desired to compute the power levels in a system consisting of a dynamic microphone with an output of -90 db (reference level 0.006 watt) which is working into a pre-amplifier with a gain of 43 db. This preamplifier is connected to a mixer with an insertion loss of -12 db and is followed by a main amplifier with a 67-db gain. This final main amplifier supplies power to a loud-speaker.

The total gain is easily arrived at by simply adding the individual gains and losses in the circuit. Thus, after leaving the microphone, the gains are

$$43 + 67 = 110 \text{ db.}$$

The losses are confined to the insertion loss -12 db. The net gain is 98 db.

The power level at the input is -90 db. This is brought up to a level of +8 db feeding the speaker. If we are experienced in the use of such units we know at once that this is a rather low speaker volume.

Example 4: The decibel is fundamentally a measure of power ratio, and, therefore, in expressing current or voltage ratios in decibels, it is necessary to write the current or voltage expressions so that they indicate the powers involved. The derivation of these formulas will become clear by studying this example.

Let P_1 equal the power input, P_2 the power output, and R_1 and R_2 the input and output impedances, respectively:

$$\text{db} = 10 \log \frac{P_2}{P_1}$$

but

$$P_1 = \frac{E_1^2}{R_1} \quad \text{and} \quad P_2 = \frac{E_2^2}{R_2}$$

Therefore,

$$\text{db} = 10 \log \frac{E_2^2/R_2}{E_1^2/R_1} = 10 \log \frac{E_2^2 R_1}{E_1^2 R_2} = 10 \log \left(\frac{E_2}{E_1} \right)^2 \times \frac{R_1}{R_2}$$

but

$$\log X^2 = 2 \log X$$

$$\text{db} = 20 \log \frac{E_2 \sqrt{R_1}}{E_1 \sqrt{R_2}}$$

Similarly,

$$\text{db} = 10 \log \frac{P_2}{P_1}$$

but

$$P_1 = I_1^2 R_1 \quad \text{and} \quad P_2 = I_2^2 R_2$$

Therefore,

$$\text{db} = 10 \log \frac{I_2^2 R_2}{I_1^2 R_1} = 20 \log \frac{I_2 \sqrt{R_2}}{I_1 \sqrt{R_1}}$$

It is seen in both of the above derivations that the impedance values cancel out only if the input and output impedances are equal, which need not necessarily be so.

When the input and output impedances are equal, the expressions may be written, $\text{db} = 20 \log E_2/E_1$ and $\text{db} = 20 \log I_2/I_1$.

In order to facilitate decibel calculations, it is customary in telephone and radio practice to assume arbitrarily the value of 0.006 watt (6 mw.) as 0 db.¹ Hence, unless some other level is indicated as 0 db, the level 0.006 watt is to be assumed. The statement that an amplifier delivers 0 db to its load would, therefore, be taken to mean that it has an output of 0.006 watt. Note that such a statement tells nothing concerning the *gain* of an amplifier, but it is, nevertheless, a convenient way of rating its output.

Example 5: It is desired to find the power output in watts of the dynamic microphone, previously mentioned as having a level of -90 db. In this example, as in Example 2, it is necessary to reverse the process employed, to find the decibels. Note carefully how this is done:

$$90 = 10 \log \frac{0.006}{X}$$

$$9 = \log \frac{0.006}{X}$$

But the antilog of 9 = 10⁹; therefore,

$$10^9 = \frac{0.006}{X} \quad \text{and} \quad X = \frac{6 \times 10^{-3}}{10^9}$$

$$X = 6 \times 10^{-12} = 6 \mu\mu\text{w.}$$

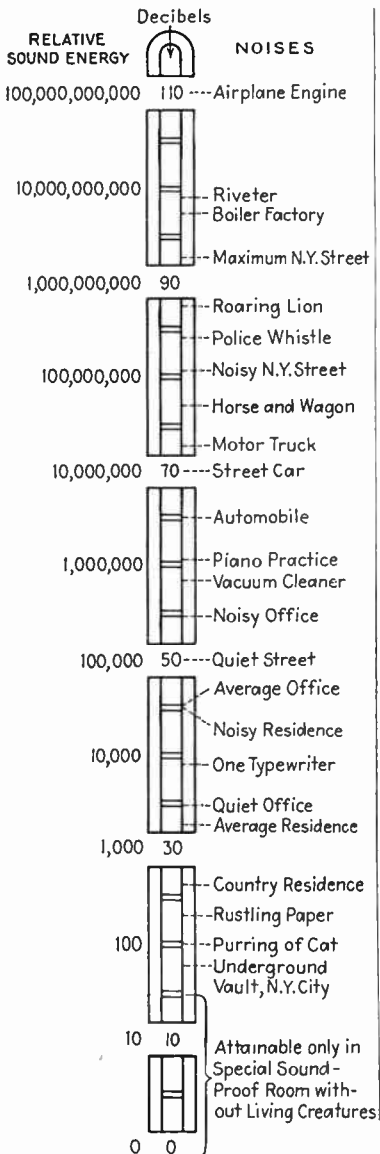


FIG. 212.—Energy level chart. (Courtesy of Electronics.)

¹ The power value which represents zero db is known as the "zero reference level."

Example 6: If the output of vacuum tubes were given directly in decibels, instead of in milliwatts, as is the present practice, it would be easier to compare their relative merits as sound-power producers.

Find the decibel rating of a -71 type tube which is listed as having an output of 790 mw.

$$\begin{aligned} 0 \text{ db} &= 0.006 \text{ watt} && \text{(reference level)} \\ \text{db} &= 10 \log 79\% \\ \log 79\% &= \log 131.6 = 2.11 \\ 10 \times 2.11 &= 21.1 \\ \text{db} &= 21.1 \end{aligned}$$

which is the output of the type -71 tube. Similarly, placing this in a list of familiar tubes the following comparative values appear.

Tube	Milliwatts	Decibels
-71A	790	21.1
-45	2,000	24.7
-50	4,600	28.8

Example 7: Connecting tubes in push pull results in an output which is 40 per cent more than double the power available from a single tube. Expressed mathematically this appears as

$$P = 2 \times W + (2 \times W \times 0.40)$$

where P is the output power in watts and W the output power in watts for a single tube.

The amount of extra power secured by placing two tubes in push pull will, of course, depend on the output from a single tube. Taking the -71 type as an example, it is seen that the output is 790 mw. or 21.1 db for a single tube. If placing the tubes in push pull only doubled the output, 1,580 mw. would result which equals 24.2 db. However, connecting the tubes in push pull raises the output to 2,212 mw. which equals 25.6 db. Thus, an extra gain of 1.4 db has been secured by placing the tubes in a push-pull circuit rather than in parallel.

Example 8: In order to have satisfactory room volume, it is necessary to deliver an approximate power of 1 watt to the loud-speaker. This equals a level of 22.2 db computed as follows (0 db = 0.006 watt):

$$\begin{aligned} \text{db} &= 10 \log \frac{1}{0.006} = 10 \log 166.6 = 10 \times 2.22 \\ \text{db} &= 22.2 \end{aligned}$$

Let us assume that it is necessary to bring up to this level the output of a microphone having an output of -100 db. It is apparent that a gain of 122.2 db is necessary, and that the output tube must be able to handle a *power level* of at least 22.2 db. Referring to the tabulation given in Example 6, it is seen that a type-45 tube would be satisfactory for this purpose.

Example 9: As the decibel is so directly related to the effect of sound on the ear, it is the ideal unit to express the fidelity rating of a transmission circuit or any of its components. For example, let it be required to calculate the fidelity of reproduction of a certain 1:1 transformer when 1 volt across the primary winding produces in the secondary at given frequencies the results which follow:

Cycles	Voltage
30	0.1
50	0.5
100 to 8,000	1.0
10,000	0.8

This information means very little unless translated into decibels. Giving the arbitrary value of 0 db to 1 volt the following results appear.

At 30 cycles,

$$\text{db} = 20 \log \frac{1}{0.1} = 20 \log 10 = 20 \times 1 = 20 \text{ db loss}$$

This might have been written

$$\text{db} = 20 \log \frac{0.1}{1} = 20 \log 0.1 = 20 \times -1 = -20 \text{ db.}$$

This, however, would involve the use of negative characteristics which are difficult to work without a great deal of practice. It is easier, as explained before, to work the problem with positive values, that is, to put the higher value as the numerator in the fraction and change the sign in the answer. It is quite apparent that the output at 0.1 volt will be less than at 1 volt, and, therefore, the answer must be a negative value to express a loss.

Using the same method of computation and taking the voltage values from the preceding tabulation, the following results are obtained. At 50 cycles the loss is 6 db (-6 db). At 10,000 cycles the loss is 1.92 db (-1.92).

Example 10: The power of a certain broadcast station is 1,000 watts and this power produces an audio output of 10 db (0 db = 0.006 watt) in a distant receiver. The power of the transmitter is raised to 5,000 watts. What is the output of the receiver under the new conditions?

Assuming that the audio-output power of the receiver varies directly as the power transmitted, then $P_r \propto P_t$, where P_r is the audio-output power and P_t the power transmitted. Then

$$\text{db} = 10 \log \frac{P_2}{P_1}$$

Therefore,

$$\text{db} = 10 \times \log \frac{5}{1} = 10 \times \log 5 = 10 \times 0.698 = 6.98 \text{ db}$$

Ans. New output in receiver is $10 + 6.98 = 16.98 \text{ db}$.

Example 11: If it is desired to compute the voltage gain of an audio-frequency amplifier such as illustrated in Fig. 146(a), for example, it will be necessary to obtain the ratio of the output voltage divided by the input voltage. The complete procedure may be obtained by the reasoning that the voltage applied to the grid of a tube is dependent upon the square root of its input and output impedance ratio. Consequently the total voltage developed in the plate circuit due to the grid swing would be

$$E_{p(\text{a.c.})} = \mu E_g \sqrt{\frac{Z_p}{Z_g}} \text{ volts} \quad (165)$$

Thus, since the impedance of the output winding, Z_o , must at all times be at least equal to the tube impedance R_p , the IR drop across this winding would be *one-half*

of the total voltage developed in the plate circuit. Hence, the voltage developed across the secondary winding connected to tube 2 would be

$$E_o = \frac{\mu E_g}{2} \sqrt{\frac{Z_i}{Z_o}} \text{ volts} \quad (166)$$

Thus the total voltage gain of any amplifier will be the ratio of the output voltage divided by the input voltage E_o/E_i , or,

$$\text{Gain} = \frac{\mu}{2} \sqrt{\frac{Z_i}{Z_o}} \text{ volts}$$

The numerical gain ratio may then readily be converted into decibels by referring to the decibel and amplification table in Sec. 238 where the letter K is used to denote the gain ratio.

As a practical example let us assume that the input transformer in Fig. 146(a) has a secondary impedance of 360,000 ohms. The tube is a -27 type with an amplification factor of 9 and a plate impedance R_p of 9,000 ohms. Now, since the load impedance is at all times assumed to be at least matched to the tube impedance, we may compute the voltage amplification gain by the ratio

$$\text{Gain} = \frac{9}{2} \sqrt{\frac{360,000}{9,000}} = 28.5 \text{ voltage ratio (approx.)}$$

Referring to Tables IV and VII we may judge that a voltage amplification of 28.5 is equivalent to a gain of slightly over 29 db.

The decibel gain of a complete amplifier system may thus be readily computed by applying the equation to each successive stage as shown.

These examples show the convenience of using the decibel to give a mental picture of the gain and output power of a transmission system. The decibel may also be used to show losses, but this particular use of the decibel will be taken up in detail when attenuation networks are studied, in the next chapter.

It is evident that a thorough understanding of the decibel and its uses must be possessed by any technician handling broadcast or public-address equipment.

As a matter of practice, the student is urged, after studying this section, to convert decibel values into equivalent power values, and vice versa. For this purpose it may be assumed that the input impedance to the hypothetical system under consideration is 50 ohms, and that the output impedance is 5,000 ohms, the problems being worked in terms of voltage or current values.

Other calculations involving the decibel will be found throughout the remainder of this chapter and also in the next chapter.

II. MICROPHONES

193. Microphones—General Considerations.—The microphone, which is also known as a *transmitter*, is a device which changes acoustical energy, consisting of variations in air pressure at audio frequencies, into corresponding electrical variations in an electrical circuit. If the changes

in the electrical circuit follow the sound impulses exactly, the microphone is said to have *perfect fidelity*.

Before dealing with microphones in particular, let us examine the general conditions which they must fulfill.

If it is desired to reproduce with absolute fidelity the sound generated by a symphonic orchestra, which contains practically all the frequencies likely to be encountered in music, the microphone would necessarily have to respond equally well to all audio variations from 30 to 15,000 cycles or even higher. It is also important that no extra frequencies not present in the original be added in the microphone. Such addition is known as *harmonic distortion*.

If all the frequencies from 60 to 8,000 cycles are present in proper proportion, the results will be very realistic and the quality of reproduction high. Indeed it will probably be quite satisfactory to the average listener if all the frequencies between 60 and 5,000 cycles are reproduced alike. In order to recreate speech satisfactorily frequencies from 60 to 5,000 cycles should be present. However, frequencies from 3,000 to 5,000 might be lost without impairing intelligibility to any great extent. The frequency range from 60 to 300 cycles affects the naturalness but not the intelligibility of the reproduction.

A microphone must be capable of reproducing great variations in sound pressure without distortion. A symphony orchestra at full volume may exert a pressure of 300 dynes per square centimeter on its diaphragm. The average studio program will produce 10 dynes per square centimeter, and probably the lowest pressure to which it is expected to respond is 1 dyne per square centimeter.

If the microphone itself generates any sound, such as a hiss or a sputter, this sound will be noticeable in the background of soft passages of music, and during dead air, unless it is at least 40 db below the weakest signal encountered.

It is desirable to locate the amplifier at some distance from the microphone. This at once makes the output impedance highly important. For maximum power transfer the impedance of the microphone and its load must match; hence high microphone impedance necessitates high load impedance. Any connecting line between microphone and amplifier will possess distributed capacity; this may be roughly considered as a condenser of equivalent value across the load.

Let us see what the effect of this distributed capacity is, by taking as an example two microphones having exactly the same characteristics, except that the impedance of one is 50 megohms, and that of the other 50 ohms. They are each connected to an individual amplifier, matching their impedance, by a microphone cable having a distributed capacity of 31 mmf. per foot, and which is 30 ft. in length, giving a total distributed capacity of 930 mmf. across the load.

The microphones deliver the same output at 50 as at 10,000 cycles. Consider the 50-ohm unit working into the 50-ohm load. If there were no distributed capacity in the cable the same energy would be delivered to its termination at all frequencies, but at 50 cycles the distributed capacity of the cable under consideration presents a reactance of about 344×10^4 ohms ($X_c = 1/2\pi fc$), and at 10,000 cycles one of about 17×10^3 ohms. If we take the energy delivered by a hypothetical non-capacitive cable as standard, the power loss occasioned at the highest frequency by the use of the considered cable will be about 0.5 per cent. At the lowest frequency it will be less than this. Thus a much longer cable than the one considered may be used with the 50-ohm microphone without noticeable effect on the frequency response of the system or too great an energy loss taking place.

Considering again, the 50-megohm microphone, if the cable possessed no distributed capacity, the energy delivered would be the same at all frequencies. Let us call this energy level 0 db. At 50 cycles the loss introduced by the 30-ft. cable having 930 mmf. capacity, will be about 24 db. As the frequency increases, the loss will increase, until, at 10,000 cycles, it will be about 70 db. With a microphone of such high impedance it can be seen that a connecting cord more than a few inches long cannot be used unless special precautions are taken against distributed capacity.

The foregoing theory applies equally well to program transmission lines and is one reason why the impedance of apparatus working into or out of such lines is kept low, if this is at all possible.

The student is urged to check the foregoing statements mathematically for the practice it will afford.

The position of the various microphones in the studio system is shown in the block diagram of studio and control-room equipment (Fig. 230).

Output levels in decibels of the various types of microphones and pick-up devices as given in this chapter are average values. As such they may not be exact when applied to devices of a specific type of manufacture as changes in design are constantly being made.

194. Carbon Microphone.—The operation of this type of microphone is based upon the phenomena involved when the pressure is varied upon a pile of carbon granules. The resistance of the pile varies directly as the pressure exerted upon it. If a diaphragm is so arranged that the audio-frequency waves impinging upon it cause a variation in pressure on a carbon pile which is carrying an otherwise steady current, variations in that current will occur. These will be a faithful reproduction of the diaphragm movement in amplitude and frequency. In order, however, to have the diaphragm movement a faithful reproduction of the audio wave, that unit must be tightly stretched to eliminate mechanical resonance at the lower frequencies. This limits its displacement to a great degree and lowers the output level otherwise obtainable. Thus, in the

ordinary carbon-type telephone transmitter, where it is desired to reproduce frequencies from 200 to 3,000 cycles with a very low degree of accuracy, the diaphragm is unstretched and the output approaches 0 db, while in the broadcast model, which is faithful in reproduction from 70 to 6,000 cycles, the output is rated at -45 db.

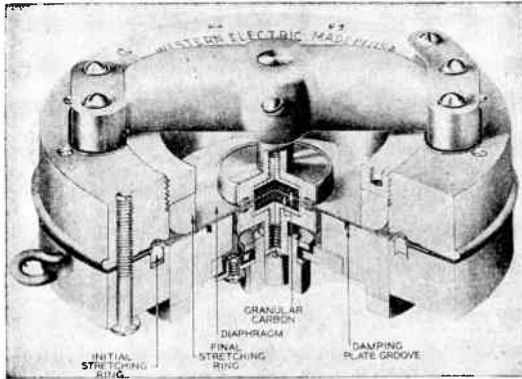


Fig. 213.—Constructional details of the carbon microphone.

The physical construction of a high-quality carbon microphone is shown in Fig. 213. It will be noted that two piles of carbon granules or buttons are employed, with the diaphragm tightly stretched between them. This design allows the microphone to be used in a push-pull circuit, with consequent advantages.

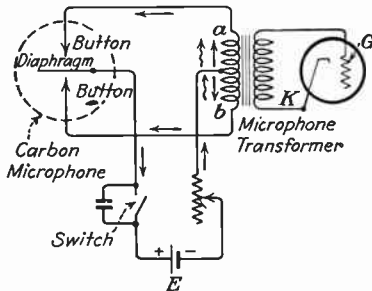


Fig. 214.—Schematic circuit of carbon microphone connected to amplifier.

We shall first consider the direct-current characteristics of the circuit as shown in Fig. 214.

In the illustration, the straight arrows show the direct-current flow. It will be noted that if both buttons offer the same resistance, these currents will be equal, and, as the lines of force generated by them oppose each other, direct-current saturation of the transformer core will be avoided. When in this state the microphone is said to balance, and for commercial models shows a current drain from 15 to 30 ma. per button. It should always be operated in this manner when possible, and, if the balance is disturbed by jarring or wind pressure, it should be restored by rotating the unit gently while the current is "off." If the current is left on, permanent damage to the microphone may result. If it proves impossible to bring the button readings to within 10 ma. of each

other, the microphone should be returned to the factory for repacking. When it is rebuilt in this manner, it is as good as new.

As the buttons possess neither inductance nor capacity, their alternating-current and direct-current resistances are the same, approximately 100 ohms per button. Thus the microphone presents a 200-ohm impedance to its input transformer and 50 ohms resistance to the direct-current supply source. The voltage at the microphone should not be over 2 volts, and, as the supply potential is usually higher than this, a rheostat is connected in series with the circuit to reduce the voltage to the proper value. Precaution must be taken to prevent current surges, which might be destructive to the delicate carbon granules, when the highly inductive circuit is broken.

A condenser connected across the battery supply switch helps to minimize this effect and greatly increases the useful life of the device.

Now let us consider the alternating-current relations in the circuit. As the diaphragm moves, the pressure on one button grows greater as that on the other grows less. Referring to the Fig. 214 let us suppose a sound wave moves the diaphragm and causes an increase in current through coil *a*. As the induced e.m.f. is always in such a direction as to oppose any change in the circuit condition, the induced e.m.f. caused by this increase will be in the direction shown by the wavy arrow beside coil *a*. At the same time, because of the diaphragm movement, the current in coil *b* is decreasing. The direction of the e.m.f. induced by this change is given by the wavy arrow beside coil *b*. No matter how the diaphragm moves, the induced e.m.f.s. will always add and so appear aiding each other in the secondary winding of the microphone transformer. The output rating of the carbon microphone is -45 db.

The carbon microphone has one very serious disadvantage; it generates a steady hiss in its output owing to minute variations in the contact resistance of the carbon granules themselves. In order to keep this hiss well below program level, the audio input to such a microphone must at all times be kept high. This is not always possible and the use of this type of microphone is therefore somewhat restricted. However, its many other advantages make it very important in its field. The use of the push-pull primary circuit helps balance out this hiss to some extent, but its greatest function is to suppress any even harmonics that may be present owing to the non-linearity of the microphone response.

The *carbon noise*, as the hiss is called, sets the limits for the lowest intensity of sound that may be picked up. Large sound intensities, such as a full orchestra, may be handled without difficulty. However, if a speaker is near enough and talking loudly enough for his breath to reach the microphone in puffs, he may easily succeed in badly unbalancing the buttons or *packing* the carbon granules together. In this condition the output is distorted and gives a characteristic overloaded sound. On rare

occasions the microphone may generate a bubbling noise. This condition is due to large and sudden changes in contact resistance and is remedied in the same manner as unbalance.

Most microphones are to some extent directional in response, this characteristic usually increasing with the frequency involved. The carbon microphone is only slightly so, although the most natural reproduction is obtained for sounds impinging at right angles to the diaphragm, and, as this angle is lessened, the higher frequencies fall off. For this reason persons with sibilant voices usually talk *across* the diaphragm. As only one side of the diaphragm is exposed to sound, the microphone responds only from the front.

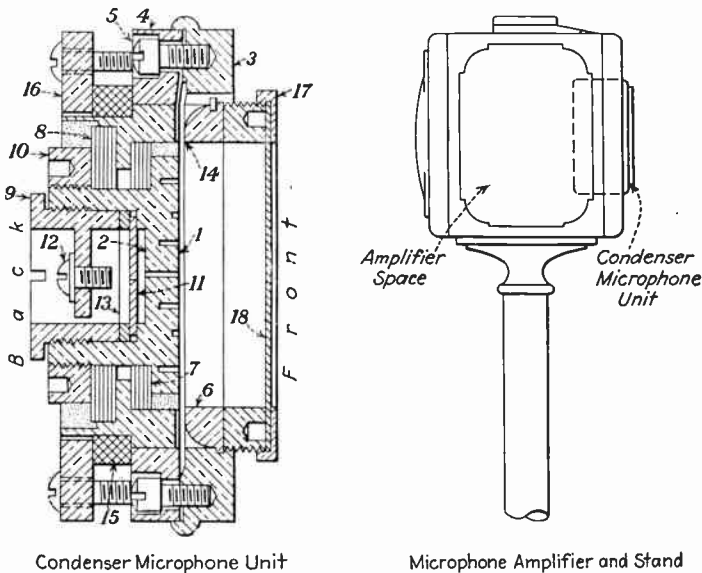


FIG. 215.—Condenser microphone.

In order to avoid the displacement of the carbon granules by mechanically transmitted shock, the microphone is always used with spring suspension. With ordinary care in handling and operation, its life may be well over a year of 12-hr. days before repacking becomes necessary.

The great advantages of the carbon microphone are its light weight, sturdy construction, and general portability. This, coupled with the fact that its output is high in comparison with other microphones, and its battery requirements low, makes it excellent for remote pick-up or portable use. Not the least responsible for its popularity is its low initial cost.

195. Condenser Microphone.—In this type of microphone, shown in Fig. 215, the diaphragm acts as one plate of a condenser. By its movement it alters the charge and, in so doing, produces a voltage drop across

a suitable resistor connected in series with it and the charging source. The voltage thus produced will be a true replica of the diaphragm movement in amplitude and frequency.

In the broadcast model, the diaphragm is tightly stretched in order to avoid mechanical resonance at undesired frequencies. The area of the diaphragm is kept small, to prevent acoustical distortion of the type referred to in the beginning of this chapter, and, despite the very close proximity of the front to the back plate, the capacity between these two units is very low.

Because of the minute movement of the diaphragm under ordinary sound pressures, and the low capacity of the condenser head, very high values of resistance and charging voltage must be employed in series with

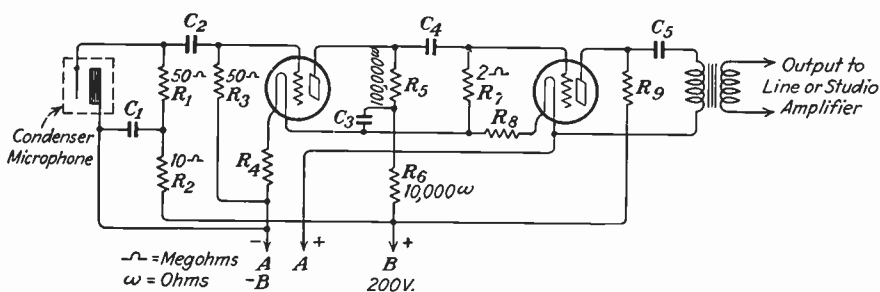


FIG. 216.—Condenser microphone and amplifier.

the device in order to secure the usable output. The resistor R_1 (Fig. 216) usually has a value of from 25 to 50 megohms and works in series with a charging potential of 200 volts. Under these conditions the output of the condenser microphone is on the order of -95 db.

The high impedance and low output of the unit make it impossible to run anything but an extremely short lead from the condenser head to its amplifier. The head and amplifier are usually mounted as a unit, and both are included under the designation *condenser microphone*.

The amplifier generally employed, and shown in Fig. 216, consists of two or three stages of low-gain tubes in a resistance-coupled circuit. As the total amplification behind the microphone output will probably be over 120 db before it reaches the listener's ear, it is of great importance that the amplifier be constructed to secure as quiet operation as possible. It must use high-quality components, especially resistors and tube sockets, and the tubes chosen should be very non-microphonic in themselves and be mounted in cushion sockets.

It should be remembered that any noise generated in this unit will be of a magnitude comparable to the program level at that point and will be amplified to the same extent. For this reason it is important to keep the microphone head free from moisture which gathers as a film across the insulation between the condenser plates and offers a variable-resistance

leakage path for the charging current. This causes loud "plopping" sounds in the output of the amplifier. Moisture on high-impedance parts of the amplifier produces the same effect. However, the tube heat generally keeps them fairly dry and free from this trouble. When condenser heads are not in use, they should be placed in a dry box, warmed by a light bulb, and kept under a polarizing potential.

It is important that the first tube in the amplifier be linear over a much greater portion of its curve than necessary to handle the normal microphone output. If this is not so, this tube will act as a detector for stray radio-frequency currents and may easily produce sounds of greater magnitude than the program. To guard against this, the amplifier and all leads must be well shielded.

A two-stage amplifier employing low μ tubes in resistance coupling has usually enough gain to bring the condenser-head output up to a point comparable with that from a carbon microphone.

Extremely low sound inputs may be handled as there is no background noise present from the head itself, such as with the carbon microphone. The highest sound intensities that can be transmitted are governed by the overloading of the microphone amplifier. As there is no gain control between the microphone and amplifier, the sound must be restricted at the input.

The frequency response of the condenser microphone is excellent from 30 to 9,000 cycles. Some of the older models have one defect which, if not corrected, has the tendency to produce a slightly mechanical-sounding reproduction. Beginning at about 500 cycles the curve gradually rises to a peak between 3,000 and 4,000 cycles. From this point it falls to reach its original level at 7,000 cycles and continues falling beyond this point to 9,000 cycles. The highest point on the curve is 8 db above the flat portion. The cause of this rise is the acoustic resonance of the air chamber formed by the recessed diaphragm. In practice it is smoothed out to some extent by the insertion of a thin layer of billiard cloth behind the diaphragm screen. This produces a consequent loss in high-frequency response but also an increase in the pleasantness of the reproduction.

The condenser microphone is perhaps the least critical, with the exception of the ribbon and crystal types, in regard to sound incidence and directional effects, although there is a loss of high frequencies as the angle of incidence departs from a right angle.

The disadvantages of the condenser microphone are lack of portability and ruggedness; the necessity for close proximity of its amplifier; its weight, large size, and necessity for a bulky battery supply. All of these factors restrict its use to studio or non-portable pick-ups. It requires constant maintenance and care, especially in damp locations.

Its advantages are high fidelity and freedom from background noise.

The output impedance of the microphone amplifier may be whatever is desired. It generally varies from 50 to 500 ohms in commercial models.

196. Dynamic Microphone.—This type of microphone, also known as the *moving-coil* type, utilizes an unstretched, non-rigid diaphragm. In it, extreme latitude of movement is achieved by utilizing a number of circular corrugations of great flexibility. Because of the large displacements thus made possible, the diaphragm response at very low frequencies is excellent. The general construction of this microphone is shown in Fig. 217.

Rigidly attached to, and free to move with, the diaphragm is a coil made up of a large number of turns of extremely thin aluminum ribbon.

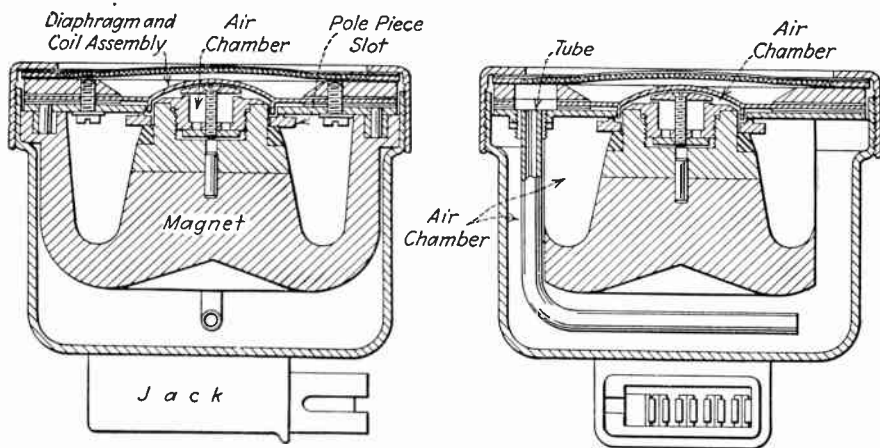


FIG. 217.—Dynamic microphone assembly.

This ribbon is insulated from itself and the diaphragm by means of a suitable varnish, which is also utilized to keep the assembly together. Because the entire movable unit is of minute weight it will respond to very high frequencies. By means of acoustical damping in the microphone assembly, excellent response is obtained over a range extending from about 20 to 9,000 cycles.

The coil is arranged, similarly to the voice coil on a dynamic speaker, to pass between the poles of a powerful *permanent* magnet, made of cobalt steel. The clearance between the coil and magnet is kept small in order to provide as intense a magnetic field cutting the coil as possible.

As the coil moves with the diaphragm in proportional response to the sound waves, it cuts the magnetic lines of force between the magnet's poles. Thus there is induced into the coil a current which, in frequency and amplitude, is a faithful replica of the diaphragm movement. The current thus produced in the coil constitutes the output of the microphone.

The impedance of the coil is low, from 25 to 50 ohms, thus it may be connected to its amplifier by means of long cables without any detriment to the frequency response of the system. However, as the output of the microphone is on the order of -85 db, the cables used must be well shielded in order to avoid picking up stray fields which might interfere with this weak signal.

As there is no background noise present, the dynamic microphone can reproduce weak signals without interference. It will reproduce equally well great sound intensities without tendency to overload. It is adaptable to low-level mixing, as a variable-attenuation network of proper design may be inserted between it and its amplifier.

In practice the dynamic microphone is found to be very rugged and may be handled (gently) while in operation without producing sounds in

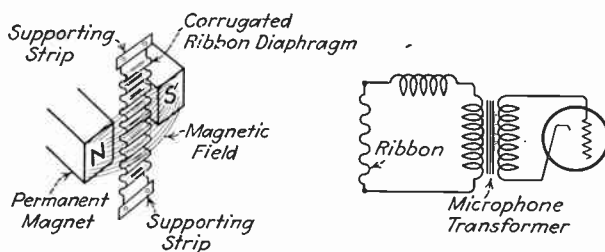


FIG. 218.—Velocity or ribbon microphone assembly and circuit.

its output. Hence, there is no necessity to mount it with spring suspension.

The great advantages of the dynamic microphone over both the carbon and condenser microphones is its dependability, independence of battery supply, and independence of constant-maintenance requirements. Added to this are the benefits of excellent frequency response, light weight, small size, and consequent portability, and its freedom from effects due to atmospheric changes. Its life is indefinitely long.

The dynamic microphone has one defect worthy of note: that is the rapid falling off of frequencies above 1,000 cycles as the angle of sound incidence departs from 90 degrees. For the most natural reproduction, then, the dynamic microphone should always directly face the sound source.

197. Velocity or Ribbon Microphone.—Unlike the types already dealt with, this particular type of microphone is actuated by the velocity component of the sound wave. Its mechanical construction is shown in Fig. 218. It consists of a very light ribbon of corrugated aluminum hung loosely between the pole pieces of a powerful, cobalt-steel, permanent magnet in such a way that the magnetic field cuts the edge of the ribbon. By allowing very small clearances, the ribbon is made free to

move back and forth between the pole pieces without touching them while, at the same time, the field strength is kept as high as possible.

As the ribbon moves in proportional response to the velocity component of the sound wave, it cuts the field between the pole pieces. In this manner there is induced into it an electric current which is an exact reproduction of the ribbon movement, in both amplitude and frequency. As the resistance of the ribbon is only a fraction of an ohm, its output is transferred immediately to the primary of an adjacent transformer, matching its impedance. This transformer is considered an integral part of the microphone assembly. Its secondary impedance may be anything desired but is usually either 250 or 500 ohms. The output obtained from the ribbon microphone is about -100 db.

Owing to its low impedance the ribbon microphone may be connected to its amplifier through long cables, which, because of the low levels involved, must be carefully shielded. The microphone may also be employed in low-level mixing circuits, though the utmost precaution must be exercised to obtain sufficiently quiet volume controls.

The frequency response is excellent from 20 to 15,000 cycles, provided that the sound source is at a greater distance than 2 ft. from the microphone. If the source be nearer, distortion will take place causing over-accentuation of the lower notes. This characteristic makes the ribbon microphone difficult to use where bad acoustic conditions must be overcome. It also helps to effect a practical decrease in output, as performers must work at a greater distance from it than they do from other microphones.

As the ribbon is acoustically shielded on two sides by the pole pieces of the magnet but is free to receive sound impulses on either of its broad surfaces, the ribbon microphone transmits sound equally well from either its front or back. As the angle of sound incidence departs from 90 degrees at either front or back, the *total response not high frequencies only*, falls, cutting off quite sharply when the sound veers around toward either side of the case. This very directional effect is sometimes valuable in helping to avoid acoustical feed back from loud-speaker to microphone when the transmitter is used for public-address work. This characteristic is often of help in correcting poor studio acoustics.

The ribbon is very delicate and must be protected from sudden strong draughts or puffs of wind. To aid in this, it is surrounded by an inner and outer screen of fine-mesh silk, attached to its magnetic assembly and outer case, respectively. These and a metallic screen help protect against mechanical injury or the accidental entry of magnetic material into the assembly.

The ribbon microphone is nearly as rugged as the dynamic type and matches it in portability, ease of maintenance, and lack of background noise.

198. Crystal Microphone. *a. General Considerations.*—This type of microphone employs a piezo-electric crystal. Such a crystal, when properly cut, if subject to mechanical stress, shows charges of opposite polarity upon its faces. The magnitude of the charges is dependent upon the displacement of the piezo-electric body. The common crystal microphone shown in Fig. 219 is made up of a number of crystal cells, arranged to aid each other and so increase the overall sensitivity of the unit. The cells are so placed that the impinging sound waves cause mechanical stress in the crystals, the resultant electrical charges constituting the output of the microphone.



FIG. 219.—
Studio type
crystal micro-
phone. (Courtesy
of The Brush
Development
Company.)

The crystal cells are so connected that while the effect of the sound wave on one cell is multiplied by the number of cells (aiding each other for sound waves), the outputs neutralize each other as to mechanical displacement. This makes it unnecessary to use spring suspension with the crystal microphone, or to protect it unduly from mechanical shock.

The superiority of Rochelle Salts over other crystals in its piezo-electric property has made its use for this type of microphone highly desirable. Although the salts are soluble in solution, rigid experiment has shown that suitable protection against moisture and humidity can be obtained by enclosing the crystal in waterproof papers and waxes. An actual test of the crystal microphone has shown that the instrument was not affected by moisture even after hours of immersion in water. The effect of temperature variation has also been reduced to a minimum by the use of very thin sections of crystal arranged in two sections or plates cemented together in such a manner that when electrodes are attached and an e.m.f. is applied, one of the plates will expand and the other plate will contract. This results in a bending effect similar to a bimetallic thermostat which practically eliminates molecular friction and saturation. The arrangement is known as a *bimorph element*. In consequence it has been found that extreme temperature variations extending from 40° below to 120°F. above zero have produced a microphone variation of only ± 2 db over the entire range.

b. Theoretical Operation.—Since it has been shown (Sec. 106) that an electrical pressure across a crystal will produce mechanical vibration, and, conversely, that a mechanical pressure will produce an electrical potential across the surfaces, the crystal may be used to convert sound-wave impulses into electrical potential variations. This may be more clearly seen by referring to Fig. 220. Here we see the homogenous structure of a Rochelle-Salts crystal in which the letters *a*, *b*, and *c* represent the crystal axes.

If the plates are cut into long slabs with faces perpendicular to the axis a and with edges at an angle of 45 degrees to the axis c , (Fig. 220(a)) they will expand and contract longitudinally in response to varying applied electrical potentials. If, therefore, a bimorph element, consisting of two salt plates cemented together and provided with proper connecting elements, is applied across a source of e.m.f., it will bend in a direction perpendicular to its major surfaces and parallel to the axis a . If, therefore, the plates are so cut that when the 45-degree diagonal of the plate, which is perpen-

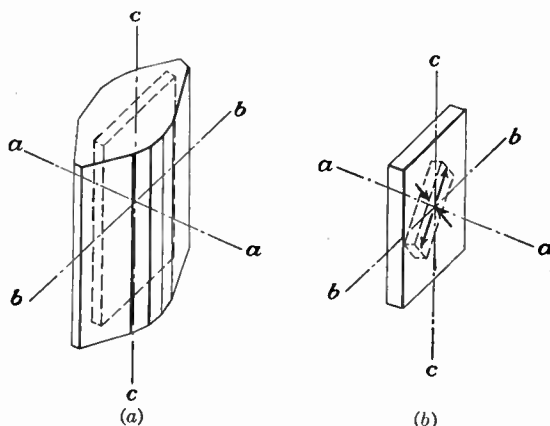


FIG. 220.—Representation of crystal axes.

dicular to the axis a , expands, the other diagonal at 90 degrees to the first will contract. Thus, if two crystal plates are made up in rectangular form, they will, upon the application of an e.m.f., assume a double curvature. Consequently, if the curvature at any instant on the 45-degree diagonal is convex, the curvature due to its 90-degree complement will be concave. Hence, if two wide elements are supported at two points in the exact center of two opposite edges, and a sound wave is impressed

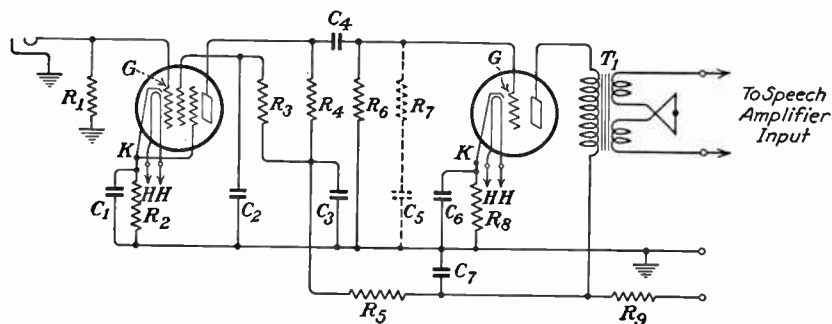


FIG. 221.—High-gain amplifier for use with crystal microphone.

upon one of the surfaces, a double-curvature movement will result, causing an e.m.f. to be developed in direct proportion to the frequency rate of the vibration.

If the microphone is connected across the input circuit of a high-gain amplifier as illustrated in Fig. 221, the operation will be as follows.

The alternating voltages developed across the bimorph element as previously described will apply an e.m.f. of an alternating character across the tube elements G , K . If the biasing resistance R_2 is properly adjusted so that the tube will operate at the straight portion of the I_p - E_c characteristic curve, the voltages developed across

the input will be symmetrically reproduced in the plate circuit in accordance with the general theory of amplification. The amplified audio-frequency variations then present in the plate circuit will develop an e.m.f. across the load-coupling resistance R_4 depending upon the value of the resistance compared with the tube resistance (R_p). These voltages are then applied through the coupling condenser C_4 to the grid of the amplifier tube. With the proper bias adjustment on this tube through R_8 , symmetrically amplified plate variations will pass through the primary winding of the output transformer T_1 . If the primary impedance of this transformer is twice the value of the plate resistance (R_p) of this tube at the lowest frequency, and if the secondary winding impedance is properly matched to the line and input impedance of the speech-amplifier unit, a flat frequency response from 50 to 10,000 cycles will be obtained. This, of course, also assumes that the plate, screen, and bias voltages are correctly adjusted for the characteristics of the tubes used.

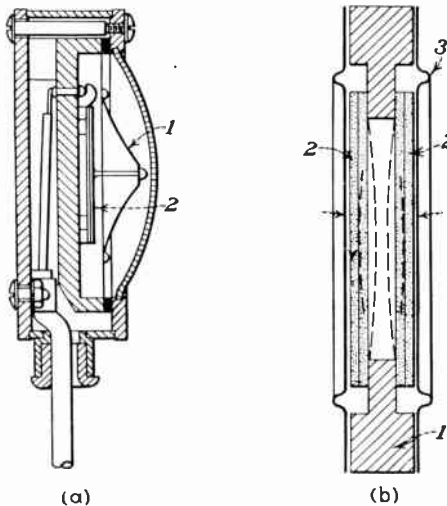


FIG. 222.—Crystal microphone assembly. (Courtesy of The Brush Development Company.)

The conventional plate-filtering system should be used throughout in order to insure an unvarying plate- and grid-voltage power supply. C_2 and R_3 form the screen-grid filter, and C_3 and R_5 the plate filter for the first tube, while C_7 and R_9 form the plate filter for the second amplifier tube. Units C_1 and C_6 are grid-biasing by-passing condensers for the two tubes. Unit R_1 is the input grid-leak resistance to enable the application of a grid bias to the tube and to prevent blocking owing to the condenser effect of the microphone elements. This resistance should be at least 5 megohms to insure a high input impedance at all frequencies to maintain the application of a steady e.m.f. across GK . Units R_7 and C_5 constitute a special compensating filter to attenuate the rise in the upper frequency spectrum near 10,000 cycles, owing to the inherent rising characteristic of the crystal microphone, in order that an entirely flat response may be obtained if desired. To attenuate frequencies below 1,000 cycles, a lower value of R_1 is used.

If the microphone is matched to the amplifier through a transformer, the impedance next to the microphone should be approximately 150,000 ohms.

c. *Commercial Models.*—The conventional types of crystal microphone employ a conical duralumin diaphragm (1) to actuate the bimorph element (2) as shown in Fig. 222(a). Two arrangements are used, however, namely, one in which the diaphragm is

rigidly connected to the crystal, and the other in which the diaphragm exerts a varying pressure on the bimorph element through a resilient intermediate member. This microphone has been improved upon to eliminate its inherent tendency to resonate at some point in the musical-frequency spectrum owing to a pressure-doubling effect caused by reflections which in turn are caused by the arrangement of the crystals in cells. This type is known as the *sound-cell* or *grille-type* microphone and is the one which apparently provides the ideal conditions for high-quality broadcasting.

The construction of a sound-cell microphone may be understood from the illustration in drawing (b). It consists of a rectangular frame of Bakelite (1) in each side of which is supported a thin Rochelle Salts crystal bimorph unit, the crystals being supported by the frame at two points (2). The space between the crystal and the frame is sealed by a flexible but air tight ring (3) which permits the crystal to vibrate freely with variations due to sound pressure. The entire cell unit is impregnated in wax at a temperature of 140°F. to maintain the elements in an airtight and moisture-proof chamber. When the entire unit is subject to sound variations, the two sides of the crystal unit vibrate in phase with each other which produces a resultant in the generation of an e.m.f. across them in direct proportions to the sound pressure, while, if the unit is subjected to mechanical vibration or shock, the voltages generated by both sides are out of phase with each other. This cancels any e.m.fs. generated by shock or vibration so that no voltages are developed across the input amplifier tube owing to these disturbances.

The physical dimension of the sound cells is one of the most important considerations in the design of the crystal microphone owing to the mechanical resonance condition involved. That is to say, if the sound cell has a mechanical resonance period somewhere in the musical-frequency spectrum, serious frequency distortion will result. This condition is prevented by designing the crystal units to have a natural resonance period above the highest frequency wave that is to be reproduced. In the commercial sound cells the crystal units are $\frac{1}{16}$ in. square and 0.020 to 0.030 in. thick, overall, making the combined bimorph unit of two crystals approximately 0.006 in. thick.

One of the exceptionally fine properties of the crystal microphone is its ability to amplify the higher spectrum of musical frequencies above 8,000 cycles. This is due to the increase in voltages developed near the mechanical resonance period of the crystal. Consequently, if the crystal has a resonant period slightly above 10,000 cycles, the frequency characteristic will rise as it approaches this point. If, therefore, a number of sound cells are connected in series, this rise may be accentuated to the point in which the overall-frequency characteristic may be brought up as high as 20,000 cycles. For commercial use, however, the sound cells are generally connected in parallel to reduce the output impedance so that reasonably long cables may be used in connecting the microphone to the vacuum-tube amplifier.

d. Electrical Characteristics.—The output e.m.f. developed by a single cell is approximately 0.125 millivolt per plate. The impedance is highly capacitively reactive and possesses a small effective series resistance, so that its impedance value near 10,000 cycles is approximately 5,000 ohms.

Owing to the fact that this microphone is purely a pressure-operated device and therefore non-directive, there is no cut-off effect at the low frequencies and consequently it will deliver a flat frequency response over the entire audio-frequency spectrum. A comparison of frequency responses with other types of microphones may be obtained by the graph in Fig. 228.

The small structure of the sound cell makes distortion due to cavity resonance, phase shifting, and wave diffraction practically negligible.

In commercial use the microphone usually consists of from 4 to 24 cells connected in series or series-parallel. If the leads from the microphone to the amplifier are of considerable length, the cells are connected in series-parallel, while, if the microphone and amplifier are self-contained, the cells are connected in series.

The output level of the commercial-type microphone with a 25-ft. cable varies between 65 and 74 db depending upon the cell connections.

Among the advantages of this type of microphone are its light weight, ruggedness, ease of maintenance, and the fact that no battery supply is necessary. Under no circumstances, however, can a gain control be inserted directly between the crystal microphone and its amplifier without a consequent loss in the low-frequency response.

The crystal microphone is entirely non-directional and may be used at any angle. It cannot be overloaded acoustically and has no inherent background noise.

199. Lapel Microphone.—Two types of lapel microphones have been successfully developed, the carbon and ribbon varieties. They are made very small and light in weight, and are meant to be worn on the coat lapel of a speaker. By means of acoustical and electrical filters their outputs are compensated to reduce the lower frequencies and to achieve passably accurate reproduction. In the carbon type the granules are packed in such

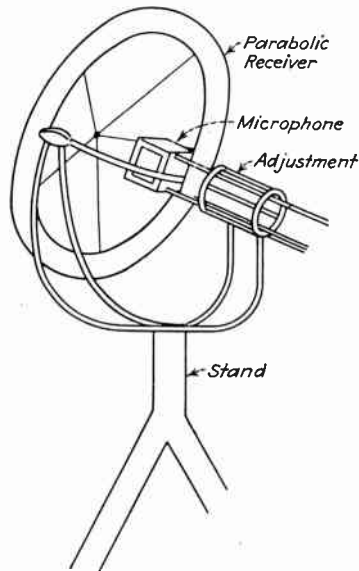


FIG. 223.—Parabolic microphone assembly.

a manner that any mechanical movement, other than that caused by the diaphragm is avoided. Lapel microphones are useful in their field but are avoided when highest quality of reproduction is desired.

200. Parabolic Microphone Receiver-reflectors.—When it is impossible to have a microphone at the source of sound, as for instance, when picking up a band marching around a football field, the parabolic microphone receiver-reflector shown in Fig. 223 is of great use. If the sound source is at some distance, the waves may be considered as being parallel without too great an error. If a microphone is placed at the focus of a parabolic receiver-reflector and the unit trained upon the distant sound, the parallel waves from that source will be converged at the microphone, and all other sound incident to the receiver-reflector will be

sidetracked past the microphone. The pick-up will then be equivalent to that from a microphone placed near the sound source.

The beam pick-up of such a receiver-reflector is quite sharp, and it is usually aimed by means of sights.

III. ELECTRICAL TRANSCRIPTION APPARATUS

201. Recordings.—Modern recording practice makes possible very faithful reproduction. In fact, if proper equipment is employed, it becomes impossible to tell by listening to the program on a radio receiver whether it is originating directly from the studio or from a recording.

A system is available whereby a sponsor may have a commercial program recorded and then distributed to various stations for broadcasting purposes. This offers many advantages over chain broadcasting, as it avoids the expensive telephone-line charges which that system entails. It also enables the program to be broadcast at the most advantageous time on each station. Because of this, part of the essential equipment of every broadcasting station is apparatus capable of accurately reproducing such program transcriptions.

Recordings are made at one of two speeds, 78 or $33\frac{1}{3}$ r.p.m. The former type of record is made in a 12-in. size and plays for 5 min. As programs usually run for at least 15 min., provision must be made to enable the operator to switch from the end of one record to the beginning of the next without the change being noticeable.

The best device for attaining this end consists of a double turn table equipped with duplicate pick-ups and arms. These arms are geared to a pointer which, by a suitable dial reading, indicates the exact location of the needle over the record. When the pick-up head is raised to clear the turn table, it locks into position and holds itself up. It is released by means of an electrically controlled catch, and, when it falls, it operates a piston which compresses a cushion of air in a piston chamber. By controlling the rate of speed at which this air is allowed to escape, the rapidity of the fall of the pick-up head can be controlled to a nicety and the needle is lowered gently into the record groove.

Switching to the right- or left-hand pick-up is governed by a special fader mounted on the turn table. In the center position both heads are "off." As the control is moved to either side, the appropriate pick-up is faded in.

In operating such equipment, the needle position at which each record starts playing, as indicated on the dial, is noted. All that is necessary then is to lift the head up and set it over the record in such a way that the indicator reads the previously "cued" number. At the end of one recording, the fader is swung over to the other side. In its passage it makes an electrical contact which releases the catch holding up the

pick-up. This then falls and, through the agency of the air cushion, is set gently upon the record, just as the control is starting to fade in its output. By this means a smoothly running program is assured. Less elaborate turn tables are provided with only a fader, the pick-ups being set manually and reliance being placed on the skill of the operator for satisfactory results.

The $33\frac{1}{3}$ -r.p.m. recordings are made on 16-in. disks which play for 15 min. As this is usually the full length of the program, no special equipment to enable rapid switching is necessary.

There are two methods of recording in use: *lateral* and *vertical* cut. In the lateral-cut record the frequency and amplitude variations of the program are carried by the undulations in the *walls* of the groove. In the vertical-cut record they are retained in the form of hills and dales in

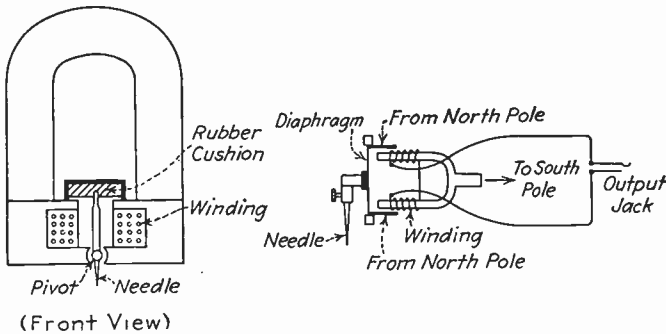


FIG. 224.—Standard magnetic type pick-up head.

the *bottom* of the groove. The latter method is not yet in general use in commercial recording of the type under consideration.

202. Pick-up Heads. *a. Standard Magnetic Type.*—In reproducing a recording by the electrical method the movement of the needle as it follows the groove variation is made to change the flux surrounding a coil in a pick-up head, the counterpart of the old phonograph reproducer. In this way an e.m.f., which, in the ideal case, corresponds in amplitude and frequency to the groove variation, is induced into the coil.

There are two distinct types of pick-ups, both employing this principle, but in a different manner. The most common type is illustrated in Fig. 224. The needle is pushed from side to side by the action of the record groove, this movement producing a corresponding variation in the armature position. With the armature in the center, the flux passes from both the upper and lower termination of one pole to the corresponding termination of the other, in equal distribution. As the armature moves to one side it creates a new path of lower magnetic reluctance from one of the upper pole pieces to the lower pole piece on the opposite side. The flux distribution is then rearranged so that the greatest number of lines

of force enter and leave the pole pieces nearest the armature. This rearrangement changes the flux distribution through the coil, inducing a corresponding e.m.f. into it. As the armature swings over to the other side, the flux is again redistributed in like manner. It will be noted that the e.m.f. induced into the coil by this change in flux will be in the opposite direction, thus completing the cycle.

In order to prevent the armature from being held to one side by the strong magnetic attraction to which it is subject upon once leaving dead center, some means must be provided for returning it to that neutral point. To provide the necessary damping and to return the armature, its top is embedded in a resilient cushion of soft rubber. The bearing at the lower pole piece is also formed by a sheet of rubber. Owing to the large proportion of mechanical rather than useful magnetic damping, entering into this construction, the efficiency is low.

The use of rubber at such a critical point of the assembly entails some important disadvantages. Because of aging, the rubber loses its resiliency in a comparatively short period of time. As this seriously affects the frequency response of the device, its dismantling for replacement of the rubbers is necessary at frequent intervals.

With this type of construction it is almost impossible to manufacture pick-ups which have identical characteristics and which will maintain them. Thus two units which originally matched each other in output may, as they age, become noticeably dissimilar. If they are installed on a double turn table, this may necessitate the replacement of either one or both of the units.

Owing to the travel of the arm across the record, there is a steady pressure against one side of the needle. In time this causes the armature to lean to one side instead of returning to the center. In order to correct this, some of the better pick-ups of this type are provided with an adjustment whereby the armature may be centered as the occasion requires.

Because of the mechanical design, it is difficult to make this type of unit respond to high frequencies. In the best examples, by resonating the armature assembly at about 3,000 cycles, the response is brought up at this point and falls from there to cut-off at about 5,000 cycles. At the low-frequency end of the curve the response is good down to about 80 cycles. Some pick-ups respond below this frequency, others have a tendency to be thrown out of the groove because of the large amplitude of the variations encountered. It must be understood that these are average curves. Individual curves will vary considerably with the age, make, and condition of the unit.

b. Oil-damped Type.—A great improvement over the standard magnetic type discussed is what is commonly called the *oil-damped pick-up*. This employs an entirely different design as illustrated in Fig. 225.

The needle is carried by a torsion-steel diaphragm which is magnetically connected to one pole of the field magnet. The other pole of the magnet terminates in a bifurcated shoe, of a high-permeability alloy, upon which are mounted the pick-up coils as shown in the illustration. By this construction the path of the magnetic flux through non-magnetic material is kept much shorter than would otherwise be possible, giving an increase in flux density available and a longer magnetic life.

With the needle in the center position, the flux density, between the torsion-steel diaphragm and both of the coil supports, is equal. As the needle is moved to one side, the diaphragm is distorted, so that it is nearer to one of the coil supports and farther from the other, causing a redistribution

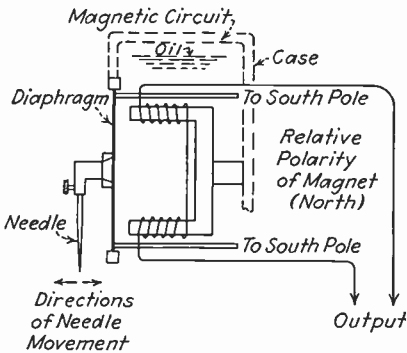


FIG. 225.—Oil-damped type magnetic pick-up assembly.

of the flux in the shoe to take place. The number of lines of force passing through the coil carried by the support nearest the diaphragm will increase, while those through its mate will decrease. The coils are connected in series so that the induced e.m.f.s. add. Thus the field about the coils varies with the needle position. It will be noted that when the diaphragm is stressed in the opposite direction, the e.m.f. induced into the coils will reverse, thus

forming a complete cycle per complete needle swing.

In order to provide greater damping than is afforded by the stress of the diaphragm alone, the assembly is housed in an oil-filled case. The oil pressing against the diaphragm increases its damping. This construction gives an output remarkably free from sudden fluctuations in the frequency curve, which starting at about 60 cycles falls gradually for about 20 db to cut-off at 5,000 cycles.

The principal advantage of this type of pick-up is its long life, uniformity in manufacture, and constant-frequency characteristics throughout its long life.

c. Piezo-astatic Type.—This type of pick-up, shown in Fig. 226, utilizes a piezo-electric bimorph crystal element to convert the mechanical motion of the needle traveling over the record into electrical energy. As this device operates by virtue of the piezo-electric ability of a crystal, its output is not subject to the effects of mechanical inertia. This characteristic gives it the name *astatic* which means that the pick-up as a body is in neutral equilibrium and, therefore, has no tendency toward any change in position, with its consequent effect on the electrical output.

The astatic nature of the device results in extremely high fidelity of reproduction which is limited only by the fidelity of the amplifier and

speaker system. The frequency response is unusually uniform and without objectionable peaks over a range extending from 30 to 8,000 cycles. Frequencies below 1,000 cycles per second may be attenuated by connecting a 0.1-megohm resistor across (parallel) the pick-up. Where the amplifier and speaker are deficient in low-frequency response a 0.5-megohm volume control is recommended.

Astatic pick-ups because of their high impedance are connected directly to the grid and ground of the amplifier tube. Where consider-

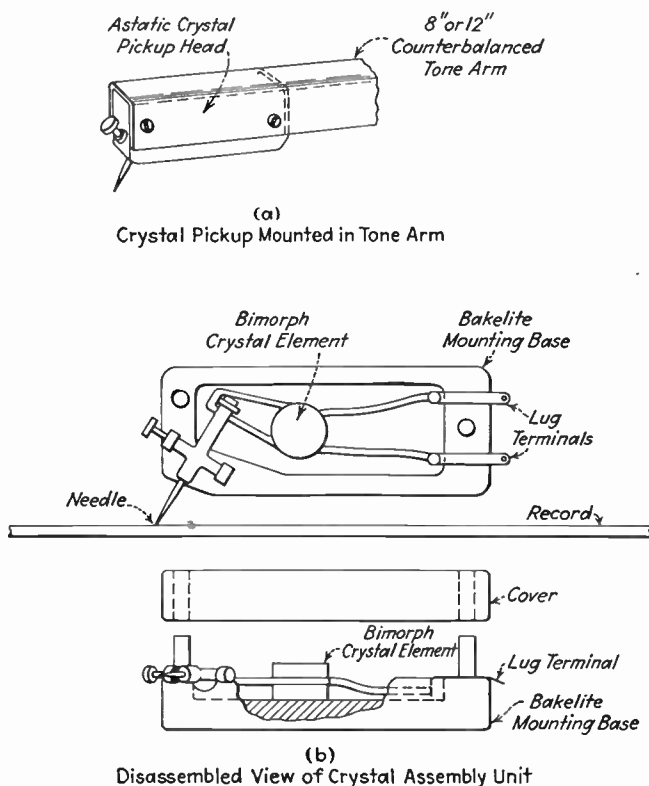


FIG. 226.—Crystal type pick-up head.

able length of cable is needed between the pick-up and the amplifier a matching transformer may be used. This transformer should have an impedance of approximately 150,000 ohms on the pick-up side.

As the crystal element is essentially a capacity, it follows that the impedance across it increases as the frequency decreases, and likewise the voltage drop increases with the impedance. The importance of this becomes apparent when it is known that commercial records are amplitude recordings in the lower frequencies, which results in strongest pick-up output where the recordings are the weakest, that is, in the bass response.

The average voltage output of the astatic pick-up from commercial records is slightly over 2 volts when using a 0.5-megohm volume control; therefore, the tube grid should be operated at from -4 to -6 volts to prevent distortion at this point.

In some installations it is necessary to mount the pick-up on sponge-rubber pads to prevent feed back from the speaker and turn-table motor. It is also important to avoid mounting the pick-up directly over the amplifier tubes or in any place where the temperature might rise above 130°F . At this high temperature the operation becomes erratic, and temperatures above this may permanently damage the element. Regular half-tone or transcription needles are used for best results.

203. Tone-arm Balance.—Tone arms on all pick-up devices should be counterbalanced in order to relieve the record of unnecessary weight,

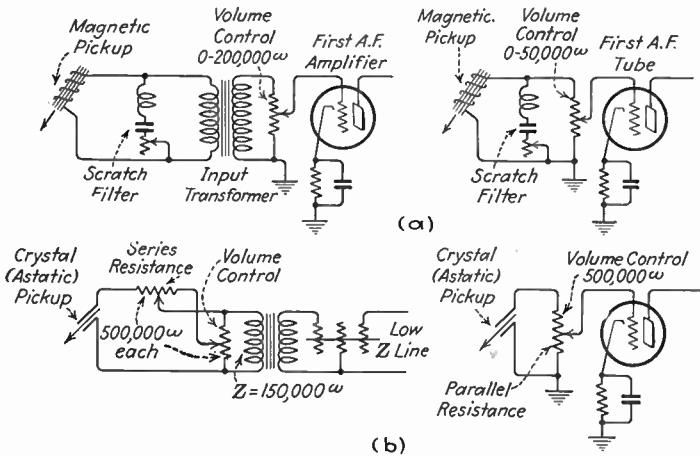


FIG. 227.—Methods of connecting the pick-up device to its amplifier showing scratch filters (a) and volume controls (b).

thereby reducing record wear and increasing the life of the pick-up head. Because of its efficient damping, the weight of the oil-damped pick-up on the groove can be made lighter than is otherwise possible without too great a sacrifice in the side-to-side inertia of the head. This effects an economy in record wear and makes this type of pick-up head very desirable for broadcast purposes.

204. Pick-up Impedance and Output.—Pick-up heads may be wound to any desired impedance. For use with broadcast equipment they are usually 50, 200, or 500 ohms, as they may then be worked into amplifiers or mixers without the necessity of changing their impedance. The usual output of a pick-up device is from -35 to -40 db. Various methods of connecting the pick-up device are shown in Fig. 227.

205. Scratch Filters.—These filters are generally connected directly across the input transformer or the pick-up device to attenuate the high

hissing frequencies due to the phonograph needle scratching. The circuit is a simple series-resonant or equalizer circuit as illustrated in the Fig. 227(a). The value of L and C is frequently tuned to resonate at a frequency of approximately 5,000 cycles and is connected in series

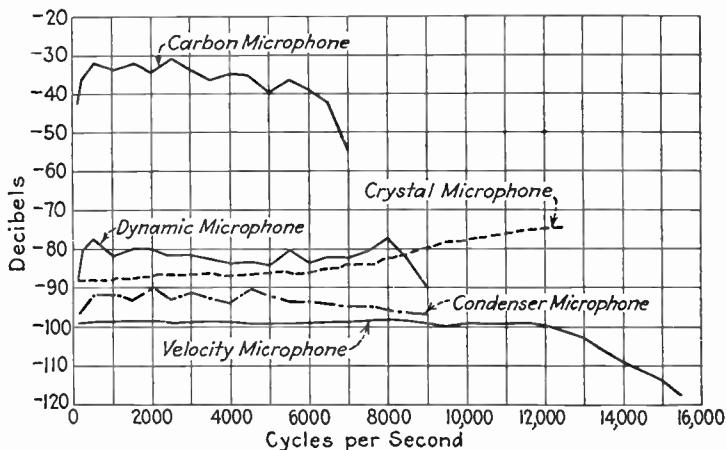


FIG. 228.—Frequency response curves of various types of microphones.

with a variable resistance to control the degree of attenuation. At the resonant frequency without the variable resistor a small band of frequencies on either side of the resonant frequency will be by-passed through the resonant circuit thereby decreasing the e.m.f. developed across it.

TABLE V.—CHARACTERISTICS OF MICROPHONES AND PICK-UPS

Type	Approximate average output, decibels	Frequency range, cycles
Microphones:		
Carbon.....	- 45	70- 6,000
Condenser.....	- 95*	30- 9,000
Dynamic.....	- 85	20- 9,000
Velocity.....	-100	20-15,000
Crystal.....	- 85	Up to 14,000
Pick-ups:		
Standard magnetic.....	- 35	80-3,000-5,000
Oil-damped magnetic.....	- 35	60- 5,000
Astatic (crystal).....	- 78	30- 8,000

* Output varies depending on amplifier employed.

Hence, if the scratch frequency is in the vicinity of 5,000 cycles, the voltage developed across LC will be small and consequently very little energy will be delivered to the input transformer unless a resistor is provided [see Fig. 243 (a)].

206. Types of Microphones and Pick-ups Compared.—In the descriptions of the various microphones and pick-ups given in this chapter, a detailed analysis of each unit and its uses was given. Table V, preceding, summarizes two of the most important factors of interest in connection with these instruments, namely, their output in decibels and the frequency range they satisfactorily cover.

Curves showing the typical response characteristics of the various types of microphones appear in Fig. 228.

NOTE: As supplementary reading to this and the next chapter the student is referred to the following reference: BALLANTINE, STUART; High Quality Broadcast Transmission and Reception, *I. R. E. Proc.*, May, 1934.

CHAPTER VIII

CONTROL-ROOM EQUIPMENT AND OPERATION

As its name implies, the control room of the studio end of the broadcasting system is the center at which converge the outputs of the microphones or other pick-up devices, and from which these outputs are distributed to the radio transmitters and to adjunct lines when chain broadcasting is involved.

I. TRANSMISSION SET-UP

207. Control-room Requirements.—A typical control-room arrangement is shown in Fig. 229. The position or sequence in the circuit of the

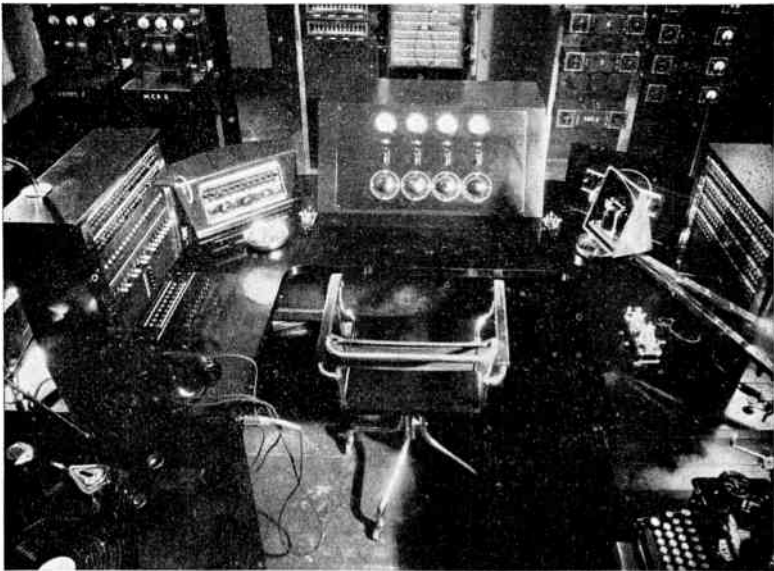


FIG. 229.—Master control desk. (*Courtesy of WCAU, Philadelphia.*)

various pieces of apparatus is shown in the block diagram of Fig. 230 which illustrates an arrangement similar to that used in the studios of the National Broadcasting Company in New York City. Both of these illustrations deserve careful attention as they tell, better than a lengthy description would do, the layout of the studio equipment.

Before taking up the control-room equipment in detail, a general description of a typical transmission system and the problems involved in its operation will be given.

In order to secure better coverage, the modern broadcast transmitter is usually situated out of town. In this way the antenna is removed from close proximity to tall steel buildings and other structures which would act as reflectors and absorbers of the emitted wave and cause a very irregularly distributed and restricted service area.

On the other hand, in order to be accessible to the required talent, it is necessary to have the studios situated at a convenient point in town.

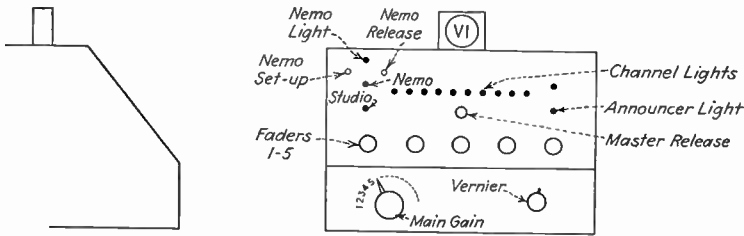
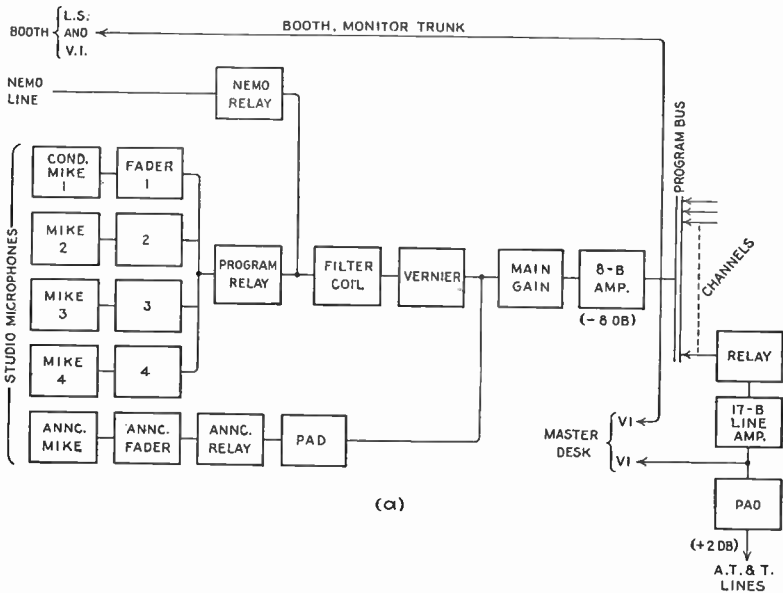


FIG. 230.—(a) Block diagram showing positions of studio and control-room equipment; (b) detail view of control panel.

Therefore, to carry the program to the distant transmitter, privately leased telephone wires which are especially prepared for radio service are required. The special treatment of these wires will be discussed later.

To avoid serious cross talk, that is, leakage from one line to another through inductance or capacity between them, the telephone company limits the normal program level to 0 db (0.006 watt) and its peak value at any instant to +2 db.

If the gain of the amplifying system were set so that the loudest passage in a program gave a line level of 0 db and modulated the transmitter 100 per cent, the program could be transmitted with absolutely faithful sound values. This, however, would not be satisfactory for several reasons. Only those persons who were situated so near the transmitter that the weakest signal broadcast was 10,000 times stronger than any extraneous noise caused by static, hum from the set itself, or any other reason would be able to receive a perfect program, free from any annoying background noise.

As the satisfactory range of a transmitter is dependent, among other things, upon its average modulation, it is desirable to keep this as high as possible. By manually controlling the gain of the amplifying system, bringing up the level of the softer passages in the program, and necessarily decreasing that of the louder passages to prevent overloading, the average modulation is increased to a point which makes it possible for thousands of additional listeners to enjoy reception unmarred by noise.

It must be pointed out that with such transmission it is impossible to have faithful sound values. For instance, in reproducing a large symphony orchestra, a volume range of 60 db may be encountered between the loudest and softest passages. This is compressed to a range not exceeding 30 or 40 db by manual control. Thus "crescendo" and "pianissimo" effects are much modified during transmission, a fact sometimes painful to the true musician, but generally unnoticed by the layman. Naturally an unskilled operator at the gain control may easily ruin an otherwise perfect program. These drawbacks are far outweighed, however, by the many advantages the system affords.

In order then, to modify the amplification and to keep the program on the line between studio and transmitter at the proper level, the transmission network must be provided with a volume-control and a level-indicating device. To vary the volume smoothly, the changes must take place in such small steps that the listener's attention will not be drawn to them. Controls having variations not greater than 1.5 to 2 db between steps prove very satisfactory, as does also the system of having a main control in steps of 3 db used in conjunction with a separate vernier control.

208. Control-room Equipment. a. Mixers.—It is often necessary to use more than one microphone to properly transmit a program, and it is essential to be able to control the output of each individual microphone at will. The device which accomplishes this is known as a *mixer*. It is usually mounted near the main gain control and the volume indicator in such a position that its controls are easily available to the studio control-room operator. Sometimes this whole assembly is mounted on what is called a *turret*, which is an elevated panel placed upon a desk in such a position that the operator may vary the controls while looking into the

studio through soundproof windows. The soundproofing of the control room from the studio is, of course, necessary so that the program may be monitored through a loud-speaker, enabling the operator to hear it as it will appear on the air.

In general there are two distinct systems of mixing, namely, *high-level* and *low-level* mixing. Each of these methods will be treated in detail later. In high-level mixing the output of the microphone is first amplified and then passed on to the mixer to be controlled. In low-level mixing, the microphone output is controlled without being previously amplified, and all the amplification takes place after the mixer output.

b. Speech Amplifiers.—Following the low-level mixer in the circuit is the main amplifier, the function of this amplifier being to bring the program up to line level. For purposes of transmission it is only necessary to raise the program level to 0 db, but for monitoring purposes a power level of at least 22 db is required for satisfactory loud-speaker operation. This may be arranged for in two ways: (1) the input of another separate amplifier, called a monitor amplifier, is bridged across the line to step the program up from line to speaker level, or (2) the main amplifier is designed to have sufficient output to feed the speaker, and its output is then stepped down, by means of an attenuation network, to 0 db to feed the line. The calculation and design of such networks will be taken up in detail later in this chapter.

In most transmission systems provision must be made to feed the program to more than one point. It may be required to supply several stations on a network, or several remote loud-speakers. In order to isolate any trouble which may occur, separate amplifiers are used for each program feed required. The inputs of these amplifiers are designed to have a very high impedance, on the order of 20,000 ohms; hence a number of them may be bridged across the output of the main amplifier, usually of 500 ohms impedance, without taking an appreciable amount of power from the circuit. As the amplifiers will transmit in one direction only, they completely prevent any trouble, such as a short circuit on a line fed by an individual amplifier, from producing any effect on the circuits taken care of by the others.

c. Volume Indicators.—When a number of lines are fed by this means, it becomes prohibitively expensive to install an individual meter-type volume indicator on each line, yet it is desirable to have an instantaneous check on the operation of each unit. This is accomplished by the use of another one-tube amplifier bridged across the line to be checked. This operates a neon tube which will break down at the higher program levels and by its characteristic glow give indication that the circuit is working. The brightness of the glow also affords the experienced operator some idea of the program levels being encountered. As the neon tube presents a variable impedance it must be isolated from the line by an amplifier.

d. Equalizers.—Besides taking programs coming from the studio, the transmission system must be capable of handling broadcasts originating at remote points, and relayed to the control room over telephone wires. Owing to the distributed capacity of the long telephone line, the high frequencies present in the program tend to become attenuated. Provision is made to correct this defect by use of filter circuits known as *equalizers*, the adjustment and design of which will also be discussed later.

e. Tone Control.—If it is desired to increase the low-frequency response in an audio-frequency amplifier system it will be necessary to attenuate the high frequencies in order to obtain the desired boosting effect. This may be accomplished by introducing a series-resonant circuit similar to that used for scratch filters or equalizers and adjusting it to about 6,000 cycles.

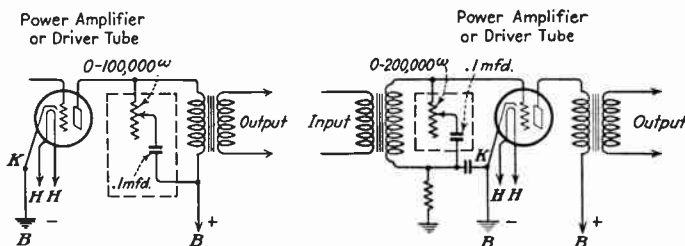


FIG. 231.—Two common methods of connecting the tone-control unit.

Two common methods of connecting tone controls are illustrated in Fig. 231. These may be connected in either the power amplifier or preceding audio-frequency stage grid or plate circuits.

These pieces of apparatus, together with the necessary relay and switching equipment, comprise the essentials of a studio layout, and the individual pieces of equipment will now be discussed in detail.

II. MIXING EQUIPMENT

209. General Requirements and Types of Mixers.—Having studied the individual characteristics of various microphones, the mixing equipment to be associated with them will now be described in detail.

The requirements for a good mixer might be listed as follows, the description being based on the low-level mixing system.

1. For maximum transfer of energy, the microphone must work into its own impedance, and the output impedance of the mixer must match the input impedance of the amplifier.

2. The impedance value presented to each microphone and to the amplifier should not vary, regardless of the position of the mixer controls.

3. The mixer should be capable of giving a smooth variation in volume from zero to maximum output. In order to accomplish this, the volume must change in steps not greater than 2 db.

4. The equipment should be capable of controlling a volume variation of at least 70 db. Half of this variation may be handled by the mixer and half by the main gain control. These are minimum requirements. Excellent results would be obtainable by the use of a mixer control having a total attenuation of 45 db, in steps of $1\frac{1}{2}$ db, used in conjunction with a main gain control of the same value.

5. The mixer should not alter the overall-frequency response of the system with which it is used.

6. The mixer should be free from any interaction between controls, the output of any microphone being absolutely independent of the control position for any other.

7. No noise should be introduced into the circuit because of changing contacts, no matter how rapidly the gain control is moved.

8. When followed by high-gain amplifiers, it is most important that the inductance of the control units of the mixer does not pick up stray fields.

9. The insertion loss (the minimum loss due to its presence in the circuit) should be as low as possible.

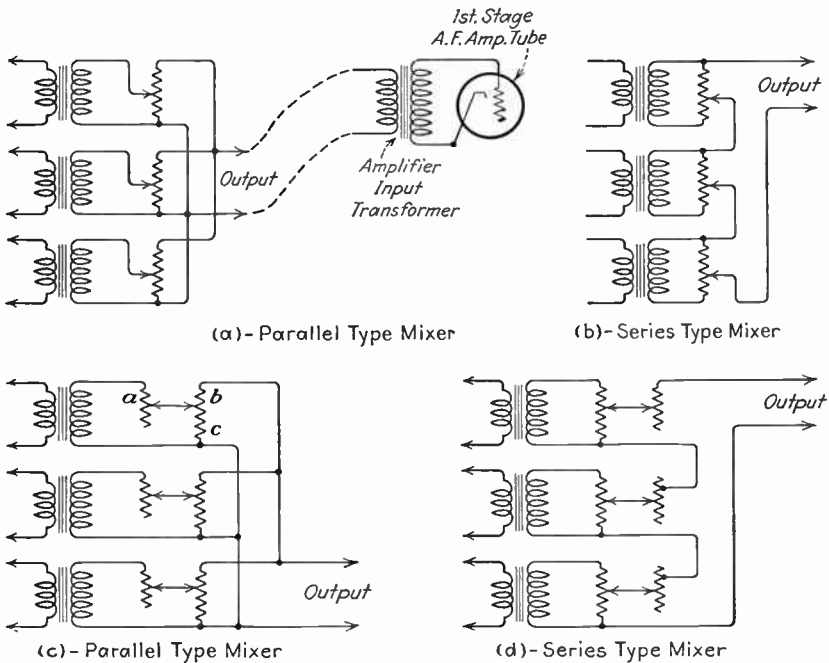


FIG. 232.—Types of mixers.

All mixers fall into one of two classes: those having the controlled outputs of the microphones in series, and those having the outputs in parallel.

The simplest forms of mixers utilize potentiometers for their controls, as illustrated in Fig. 232(a) and (b). An inspection of these figures will reveal that our second condition for a good mixer, that of constant-impedance values, cannot possibly be fulfilled by either of these examples.

This immediately obviates the carrying out of conditions 1, 5, and 6. Such mixers are seldom used nowadays.

An improvement on these simple circuits is shown in Fig. 232(c) and (d). In order to keep the input and output impedance of the control unit as nearly constant as is possible, an automatic compensating resistance is used. Movement of the control knob on this type of attenuation network causes the contact arm to slide over and connect two separate resistors. These networks are comparable to variable T pads (see Sec. 238, Attenuation Networks) as shown in Fig. 233. The loss in the circuit is occasioned by the shunting effect of resistor c . As the value of this resistor is decreased, those of resistors a and b are automatically increased. Similarly, if c be increased, a and b are decreased. With a well-designed network of this type, it is possible to have the maximum impedance variation not greater than 25 per cent. In Fig. 232(c) the first control unit has had its resistance components marked a , b , and c , in order to bring out clearly its relation to the illustrated T pad in Fig. 233.

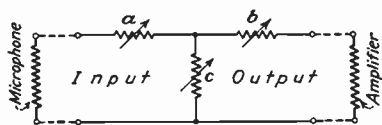


Fig. 233.—Conventional T -type pad.

The parallel type of mixer has gone out of favor, and a rough analysis of both types will bring out one of the reasons. Consider a four-microphone series mixer, working into a matching resistive load. Suppose a variation of 50 per cent takes place in the output resistance of one control, this position represents only 25 per cent of the mixer impedance, and the total variation will be only one-eighth of the whole, or $12\frac{1}{2}$ per cent. The same condition in a parallel mixer would produce a 20 per cent change in total impedance. If the mixers are working into an inductive load, such as an unloaded input transformer to an amplifier, the advantages of the series type of control will be still more apparent.

In Fig. 234 is illustrated a great improvement in mixer controls. Essentially it consists of a number of T pads in series. It will be noted that the taps for the control arm are arranged in such a way that the device does not always act as a true T pad; also that the pad termination will in most cases be improper. With a multitude of sections, however, the impedance variation caused in this type is too slight to create any audible defect in the mixer action. Such attenuators are known as *ladder networks*. They make exceptionally fine controls and may be used in either series or parallel types of mixers with equal success. As no impedance changes of any note take place in the circuit, condition 5 in our list is fulfilled, and changing the position of any one control will have no effect upon the output of any other. This effect, when present, is one of the most annoying a mixer can possess.

210. Stray Fields and Noises.—In order not to affect the frequency response of the system with which they are being used, the controls should

have their resistance units so wound or arranged that they are non-inductive and non-capacitive. Another advantage to be gained by following this practice is the elimination of trouble due to the picking up of stray fields.

At the comparatively high program levels present in the circuit when using carbon microphones or high-level mixing trouble due to stray fields is not so apt to occur. Their importance increases as the program level in the attenuating unit is decreased.

As an example, suppose a mixer control is used with a condenser microphone having an output of -35 db. A stray field from a power transformer cuts the control and induces into it a 60-cycle signal having a power level of -105 db. The alternating-current signal will be so far below program level as to be unobjectionable. Now, if the set-up be changed to employ velocity microphones, having an output of

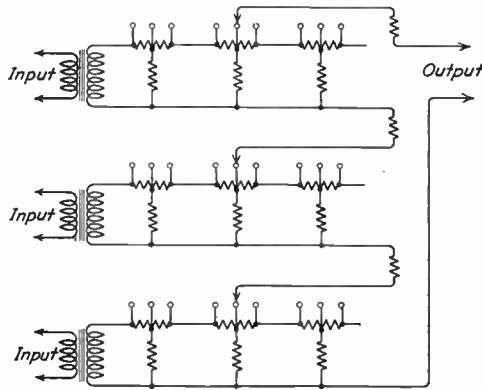


FIG. 234.—An improved type of mixer control.

-100 db, and the same control be used for low-level mixing, the 60-cycle hum in the amplifier output will be only 5 db below the program level. If a completely non-inductive control is substituted the hum will disappear.

When mixers are used to handle directly the low outputs given by dynamic- and velocity-type microphones, the tiny changes in e.m.f. generated by shifting the contacts of the volume controls, due to thermocouple effects, contact-resistance variations, and so forth, become very important. Because of this fact, there are very few controls available which will prove satisfactory for use in this manner with these microphones. Those which may be used are very carefully designed and built, and are necessarily expensive.

In order to keep controls as noiseless as possible they must be kept free from dust and dirt, cleaned frequently with carbon tetrachloride and protected from oxidation by a thin coating of watch oil or vaseline. Some types of contact noises are caused by the speed with which the arm leaves the stud. If the control employs a detent (a device for semi-permanently locking the control arm into position over each stud, usually

by means of a spring arrangement) this situation can sometimes be improved by its removal. This has the effect of converting the sweep of the control arm from a series of rapid jumps into smooth variations. The average speed of the sweep over the studs is unaffected, but the individual speed of the arm arrival and departure from each stud is considerably lessened.

When high-level mixing is used, the requirements placed on the volume control are much less severe. If the level is always kept at least 40 db higher than the disturbances introduced in the circuit by the control, the performance will be excellent. As this type of mixing requires a separate amplifier for each microphone, its convenience is somewhat offset by its expense. Where expense and maintenance are not items, it is probably the most satisfactory system.

In any circuit employing very high gains, such as a studio amplifier for use with dynamic microphones (120-db gain), very small disturbances at the input become very objectionable. Thus it is found in practice that under certain conditions, though thoroughly shielded, the cables running from the dynamic microphones to the mixer, pick up static charges. If low-level mixing is employed, with the microphone switch located ahead of the volume control, this static discharge may result in a loud "plop" in the circuit every time a microphone is cut in, even though the mixer-gain control be in the "off" position.

There are three fundamental ways of avoiding these disturbances, the first and simplest method being not to use a switch to break the microphone circuit, but to turn it off and on using only the mixer control. Aside from the technical aspect, this method has much to recommend it and is often found in practice. Its great advantage is in the operating routine. It is impossible to turn a microphone on without fading it in. Thus, if mistakes are made and the microphone is turned on a moment too late, the performer having already started, the mistake will not be nearly so noticeable as it would be if a switch were thrown in, as the change will not take place instantly but will occur in a more gradual and less attention-compelling manner.

The second method is to use a switch which automatically grounds both sides of the microphone cable when it is cut out of the circuit. If the mixer is kept at ground potential as well, no static discharge can take place when the microphone is again connected to it.

The third method is to employ high-level mixing and locate the switches in the output of the individual microphone amplifiers. In this position the seriousness of the discharge is minimized because the program level has been raised, and because the amplifier is usually located near the mixer and the leads between the two will be short.

There is another possibility for disturbance in the circuit under these conditions. If the primary of the amplifier-output transformer is above ground potential, and the static shielding between windings is not perfect, a static discharge and consequent plop will occur every time the secondary is switched into the mixer circuit. This may sometimes be avoided by parallel feeding the transformer and grounding the primary, or by employing a switching device, such as already described, which keeps the secondary at ground potential. The latter method is not always permanently satisfactory, owing to the difficulty of maintaining fine contact adjustments.

211. Insertion Loss.¹—The insertion loss caused by any item in a circuit is the loss introduced in the circuit by the use of that particular item. For example, if a variable pad (in this example a mixer) is introduced into a circuit, and this pad causes a loss of say 2 db, this loss is referred to as the minimum *insertion loss* of the pad.

Perhaps the best manner in which this condition may be fundamentally illustrated is to break down the component parts of a mixer and its associated circuits into their equivalent fundamental circuits. Let us visualize a mixer unit as a T -type pad, as in Fig. 244(b). A microphone is connected across input of the pad (Z_1), and the load is connected across the output (Z_2). This load (Z_2) represents the primary winding of the amplifier input transformer in series with a number of series pads, as in Fig. 232(d), which represents a number of additional microphone pads in the circuit. If we take, for example, a four-microphone mixer circuit (including the microphone and amplifier input transformer) and break it down into its fundamental circuit equivalents, we shall find it to appear as follows:

The microphone is visualized as an alternating-current generator looking into its mixer, which is represented for simplicity as a variable T -type pad. The output of the pad is visualized as looking into a load consisting of three resistances and an inductance connected in series. The resistances represent the three other microphone mixer pads and the inductance the primary winding of the amplifier input transformer. Now, if we assume the microphones and the pads to have a resistance of 50 ohms each, and the amplifier input transformer to have a resistance of 200 ohms, we have a picture of a simple electrical circuit equivalent. In other words, we picture a load of 350 ohms tied across the output of the variable T pad.

Now, if we desire the input and output resistance of the pad to be 50 ohms (50 ohms looking in and 50 ohms looking out), we compute a value (see Sec. 238) of approximately 40 ohms for the horizontal branches and 10 ohms for the vertical branch of the T pad. By referring again to Sec. 238 we find this amounts to a loss of approximately 20 db. In other words, by applying the rules of pad analysis, this pad when properly terminated will look like 49 ohms to the microphone, and will introduce an actual loss of 19.91 db in the circuit.

We are interested, however, in determining the loss of power by the time it reaches the mixer termination. Once more applying the rules of pad analysis, we find the power reaching the 200-ohm load to be 25.91 db below that leaving the microphone. Since 19.91 db of this loss has been occasioned by the individual attenuator setting, and the mixer loss is 25.91 db, the use of a mixer rather than a single attenuator has added a further loss of 6 db at the 20-db setting of the microphone attenuators.

If the minimum insertion loss of a mixer circuit were desired, the same procedure would be followed but the microphone pad would be calculated at its minimum setting.

The above calculations were based upon a hypothetical microphone output value of 50 volts for simplicity.

212. Special Mixing Problems.—When mixing carbon microphones other conditions must be considered in addition to those already men-

¹ Readers unfamiliar with pad calculations should study Sec. 238, p. 384, before this section.

tioned. The mixer must supply a means of measuring the individual button current without breaking the circuit. If the direct-current supply to the microphone is interrupted, either by turning it off completely, or momentarily by the insertion of a meter plug, the resultant surge of current, due to the high inductance of the circuit, will cause an extremely loud popping sound in the output.

A method whereby a meter may be inserted into the circuit without disturbing it unduly is illustrated in Fig. 235(a). The jack shown is connected in series with one of the microphone buttons. When the meter plug is inserted it lifts the previously closed contact, and the current is forced to flow through the meter. The resistance R shown should be equal to the meter resistance, thus assuring that the current in the circuit are the same.

Jacks may be avoided entirely by the use of meters cut in by switches, the circuit, however, becoming somewhat more complicated. By means of such arrangements transient currents and their consequent noises are avoided. As the carbon microphone is fed through an inductive load, a condenser (usually about 1 mf.) should be placed across the current-breaking switch in order to minimize current surges when opening the circuit. These surges might otherwise prove destructive to the delicate carbon granules. Even with such an arrangement the direct current to the microphone cannot be cut off while its transformer secondary is connected to the mixing circuit without producing a loud noise.

When it is desired to cut off the current to an individual microphone, it is first necessary to disconnect the secondary of the microphone transformer from the mixing circuit, and, as no surges result from this disconnection, it will not produce noise in the system. In order to keep the impedance relations in the circuit the same, an equivalent resistance should automatically be connected in its place. The microphone current may now be cut off without causing a disturbance in the line, *provided the internal resistance of the current source is low*. If this resistance is not low, the change in current through it will cause a change in the voltage drop across it, producing a sudden rise in the voltage applied to the remaining microphones with a consequent current surge and its resultant noise.

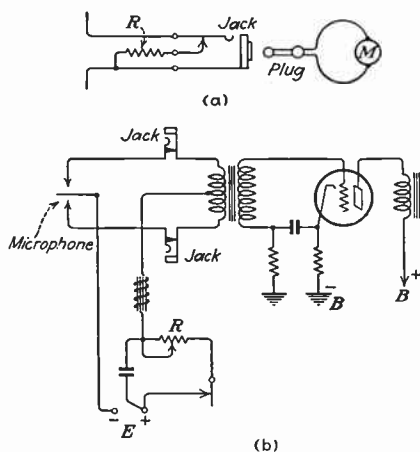


FIG. 235.—(a) Method of inserting meter in mixer circuit. (b) Method of isolating microphone circuit a. f. filter.

There is another trouble which sometimes appears when using carbon-microphone mixers. If the source of current has a fairly high resistance, and if the individual microphones are not exactly in balance (they very seldom are), the audio-frequency voltage generated across the internal resistance of the source, owing to the varying current through one microphone, may be impressed upon another microphone and appear in its output. In this way, although its mixer control is in the "off" position, the current variations in a microphone may still appear in the output of the mixer. In order to avoid this condition it is sometimes necessary to isolate each microphone-current circuit by means of an audio-frequency filter consisting of choke coil and condenser, used as shown in Fig. 235(b). The method of connecting the rheostat should be noted. This is good practice for the reason that, if the moving arm fails to make contact, the circuit is not opened thereby, but instead the total rheostat resistance R appears in the circuit.

213. Adjunct (NEMO) and Pick-up Mixing.—The program fed into the mixer does not always come directly from a microphone but may be fed in from a phonograph pick-up, or from a remote point by telephone line. In the former method the output impedance of the pick-up head must be matched to the input impedance of the mixer, usually by means of an impedance matching the transformer, or, if there is ample output, by means of an impedance-changing attenuation network. In any event, the output of the pick-up should be attenuated to approximate that from the microphone before being introduced into the mixer.

The impedance of a telephone line is generally taken to be 500 ohms. Before connecting it to the mixer it should be passed through a repeat coil designed to terminate it and which is provided with an efficient electrostatic shield between its windings. The output of this coil should be sent through an impedance-changing network in order to match the impedance of the mixer. The attenuation is necessary as the program starts on the line at zero level, and, though this level has been reduced by transmission and equalization losses before it reaches the repeat coil, it is still too high for direct introduction to the mixer circuit.

In practice it is found that for systems designed to handle dynamic microphones, or others having comparable outputs, an H pad (see Sec. 238) having an attenuation of about 40 db works well. If the mixer is used with carbon or similar microphones, an attenuation of 20 db is generally used. Regardless of conditions, provision should always be made for the insertion of extra networks in the circuit, should they be required.

The proper termination of remote lines will be discussed in detail under Equalizers and Equalizing in Sec. 235.

After these modifications of their outputs have been made, the program from either the remote-control point (adjunct) or the phono-

graph pick-up may be handled by the mixer in exactly the same manner as the output of a microphone.

III. STUDIO AMPLIFIERS

214. Special Precautions against Disturbances.—Modern studio-amplification systems are designed to work with dynamic, velocity-type, and crystal microphones. These microphones have a very minute output and require tremendous amplification following them. With the introduction of such equipment many problems which are of a minor nature when encountered in the comparatively low-gain amplifiers following a carbon or condenser-type microphone now become of major importance.

When it is realized that the power delivered by the microphone to the input of the amplification system may be less than 6×10^{-13} watt, and that any power level impressed upon that input may, by the time it leaves the transmitter, be amplified 10^{12} times, it becomes apparent that even extremely small disturbances in the input circuits of the system will be of a magnitude comparable to the program level and be amplified to about the same extent. It follows then that every precaution must be employed to keep such disturbances from entering the circuit.

To prevent the picking up of stray electrostatic and magnetic fields by the microphone cable, it must be thoroughly shielded. The connecting plugs and sockets at the cable ends are also of importance. Their construction must be such as to insure absolute and firm contact at all times as even the slightest variation in contact resistance will produce serious noises in the circuit.

The choice of input tubes in the amplification system is very limited as they must be especially quiet in operation. Most so-called *non-microphonic* tubes fail to justify that appellation or prove satisfactory under such stringent requirements. In order to protect them from mechanically transmitted shock the tubes are mounted in cushioned sockets. If a loud-speaker is operated near by, it is sometimes necessary to insulate the first tubes from acoustical shock by surrounding them with a blanket of cotton batting. If this is not done, the acoustical feed back from the loud-speaker may set the tube vibrating at its mechanically resonant frequency. This displaces the tube elements changing their spacing at the same frequency. The change in spacing will, of course, result in corresponding variations in plate current. These are amplified and impressed upon the loud-speaker from which they issue as sound waves and reinforce the vibration of the tube already taking place. In this way a note may be built up to great volumes in the same way that a microphonic detector tube subject to vibration from a loud-speaker sometimes causes a radio set to "howl."

If snap switches or gain controls employing detents are mounted on the same rack as the input amplifiers, the vibration they occasion may prove objectionable. This sometimes necessitates the removal of the

detent from the control. When alternating current is used to heat the emitter (cathode), the problem of preventing hum from entering the circuit is very difficult. This problem will be discussed in detail under Alternating-current-operated Equipment in Part IV of this chapter.

The tube sockets too are very important, for, if the contacts are not firm at all times, or if they tend to oxidize, troublesome noises are introduced into the system. They should be so constructed that, when the tube is inserted, the prongs "wipe" the contact arms, insuring a clean connection between the two. To prevent oxidization, gold-plated contacts are sometimes provided on both socket and tube.

Any disturbance introduced by the input transformer is also greatly amplified. Thus, due to the movement of core laminations, we find microphonics present in transformers. These may cause serious disturbances if the core is subject to vibration or shock. Much more important, however, is the picking up of stray alternating-current fields. To avoid this the transformer inductance is kept as low as possible, and individual magnetic shields of heavy permalloy or soft iron are employed to enclose each unit. Sometimes the transformers and choke coils in the apparatus are oriented for least coupling to the field, wherefore their arrangement in the amplifier should not be disturbed by the operator without due forethought. The low-level (input) amplifiers are mounted as far away from any alternating-current field as possible. This source of hum is perhaps the most difficult of all to avoid.

No shielding can be perfect, thus added precautions must be taken to prevent entry of undesired currents through input lines such as microphone cables.

Whenever possible, the input transformer is center-tapped and this point is grounded. This allows any currents induced into the line to flow through the primary windings to ground in such a way as to neutralize themselves.

It is important then to keep the resistance of each conductor in the cable of any transmission line at the same value. If the line is unbalanced, that is, if the conductor resistance is not equal, the induced currents will be unequal and will fail to balance out. It is for this reason that any network inserted in such a circuit should be of symmetrical construction, presenting the same resistance to each leg of the cable. How this is accomplished is taken up under Attenuation Networks, Sec. 238.

If the microphones are used in a mixing circuit it is not always possible to ground the center tap of the amplifier input transformer. The legs of the circuit may not be balanced to ground and the effect of any currents induced into the individual lines may be increased instead of diminished.

Some of the methods used to eliminate disturbances, due to electrostatic fields and currents, have been taken up in Sec. 210. In the

amplifier itself any transformer connected to a line, subject to such disturbances, should have an electrostatic shield of copper between its windings.

On dry, cold days, especially if the studios employ heavy carpets, persons may easily pick up and store large static charges, caused by friction of their feet with the floor. When they approach a grounded object or another large body they are apt to lose these charges in the form of sparks. Such discharges release a large amount of radiated energy. Hence, when using amplifiers with gains of 100 db or more, it is not unusual to find that the mere spark between two people in the studio or control room, though they are physically separated from any apparatus, produces audible clicks and plops in the equipment.

To minimize the effect of radio-frequency discharges on the apparatus, it must be most thoroughly shielded, and, in order to prevent radio-frequency-detector action, the first few tubes in the circuit must be capable of operating linearly over a much greater extent of their curve than is employed by the program. Even with these precautions, if the discharges are severe, it may prove almost impossible to sufficiently minimize their effects.

The higher the extent of the frequency range reproduced, the more natural and lifelike will be the reproduction. Modern broadcast stations faithfully transmit frequencies up to 10,000 cycles, and newer installations have equipment capable of reproducing up to 15,000 cycles. The radio receiver has not yet reached such perfection. Even the best of broadcast receivers will not reproduce to any extent over 5,000 cycles, and most sets start falling off very rapidly in their response characteristics at 3,000 cycles. However, it is possible that, in response to listeners' demands, receivers having a much wider frequency-response range may be expected in the future.

215. Harmonic and Amplitude Distortion.—In addition to the localized distortion sometimes caused in audio amplifiers (see Sec. 140) there are other types of distortion likely to be encountered in studio systems owing to the wide range of frequencies covered by these equipments. As the frequency range of the apparatus is extended many new problems arise. One of these is harmonic distortion originating in component parts of the system. By way of illustration, let us study the following situation.

Suppose a microphone capable of responding to frequencies from 60 to 6,000 cycles be used with equipment cutting off at 5,000 cycles. The quality of reproduction would be considered as being good, and it would be pleasant to listen to. The microphone has strong harmonic components, however, at fundamental frequencies higher than 3,000 cycles, but, as its lowest harmonic will then be 6,000 cycles, these strong-harmonic frequencies will not be reproduced by the amplifier (it cuts off at 5,000 cycles) and will not appear in the output. This same microphone when connected to an amplifier which is efficient up to 15,000 cycles, might give very unpleasant

reproduction, as with such equipments these harmonic frequencies, ranging from 6,000 to 15,000 cycles will be amplified to their true extent making audible those frequencies (harmonics) which heretofore had been filtered out by the poorer amplifier.

In the same manner harmonic distortion occasioned by tube overload in amplifiers is not nearly so objectionable with equipment of restricted frequency range as it is in apparatus of higher fidelity characteristics.

To understand the importance of harmonic addition or suppression in broadcasting, the student must have some fundamental knowledge of the reproduction of musical tones. Suppose that we hear middle C played upon a violin, a flute and a French horn, successively. One recognizes it as the same note in each instance, and this is because that particular note has a frequency of 256 cycles per second regardless of its source. Yet there is no difficulty in telling which instrument is playing at any time. The difference in the quality or timbre of the note lies entirely in the number of harmonics present and their proportion to the fundamental tone as shown in Table VI. It is this that enables us to distinguish between the instruments. Thus, if the harmonic content of a note is altered in any way, it changes the timbre of the tone.

The difference in the tone achieved by a great violinist and that of a sawing fiddle player lies wholly in the different harmonic contents of the same fundamental note. If the instrument adds or subtracts harmonics it is then impossible faithfully to reproduce tone values. In this way a naturally pleasing note may be given harsh and unpleasant characteristics when reproduced. Distortion of this type must be guarded against very carefully in amplifying apparatus, as it does not show up to any extent in the usual frequency-response curve. When present to only a small extent, harmonic distortion is not particularly noticeable to the ear

TABLE VI.—RELATIVE STRENGTH OF PRIME NOTE AND HARMONIC COMPONENTS

	Violin; energy, per cent	Flute; energy, per cent	French horn; energy, per cent	Violin note reproduced in ordinary tele- phone receiver (note distor- tion); energy, per cent
Prime*.....	60	13	2	8
Harmonics:				
Second.....	8	40	10	50
Third.....	20	10	50	8
Fourth.....	10	20	15	2
Fifth.....	2	5	5	
Sixth.....	..	2	2	
Total.....	100	90	84	68

* Middle C.

unless a direct comparison can be made between the original and transmitted sounds, or unless one is very familiar with the tones involved. If direct comparison is made between apparatus having some harmonic addition present and other equipment free from this defect, the output of the latter will appear to be clearer and cleaner than the former.

Tube Harmonics.—Let us consider harmonic distortion which is sometimes generated by the tube itself. Any periodically recurring wave shape, which is not a sine wave, can by analysis be shown to be made up of a number of sine waves having different amplitudes and frequencies.

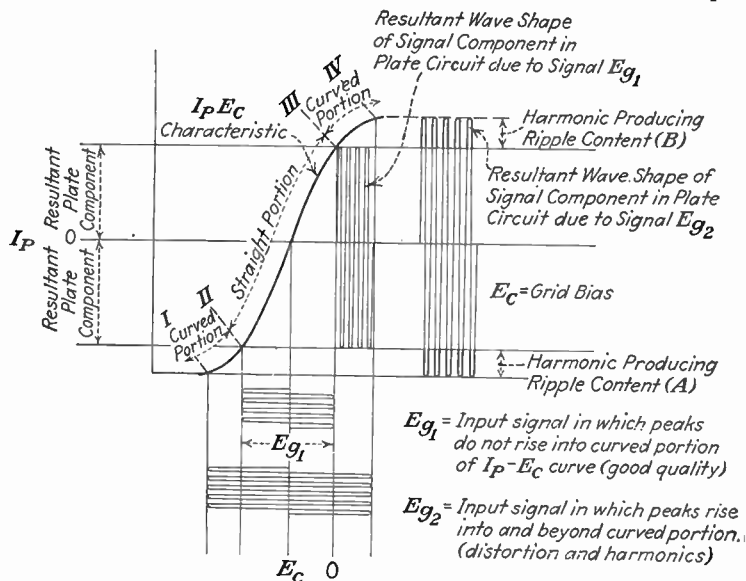


FIG. 236.—Analysis of harmonic distortion.

This is an important statement to the student and he should regard it as such. If, then, by passing a sine wave through any piece of equipment we even slightly alter its shape so that the upper and lower half cycles do not correspond in amplitude, the wave so produced will no longer be a single frequency but will be made up of the original frequency and other added harmonic frequencies of different amplitudes. The greater the distortion of the wave shape, the greater the percentage of harmonics present.

If the grid-voltage swing of a tube is limited so that the plate current is never driven off the straight-line portion of the characteristic curve (see Fig. 236), the resultant change in plate current will be an exact reproduction, in so far as wave shape is concerned, of the incoming signal on the grid. Under such conditions no frequencies that are not present in the input are added to the output, and the tube is free from amplitude distortion. This is known as a class A amplifier.

By inspection of the curves it will be noted that if the grid voltage (E_{g2}) swings the plate current down to such an extent that the curved portion of the characteristic is involved, the resultant plate current (I_p) over this part of the curve (points I and II) will be *greater* than if it were a straight line between these points. Such a grid swing will also drive the plate current (I_p) up on the upper curve at the next half-cycle (points III and IV). Here it will be noted that the resultant plate current I_p will be *less* than would be given by a straight line up to this point IV.

Hence the peak of the negative swing will be amplified to a different extent from that of the positive swing, resulting in a distortion of the wave shape and the addition of harmonic frequencies which that entails.

In practice such distortion can be shown visually by the insertion of a direct-current meter in the plate circuit of the tube involved. A visual indication is much more reliable in every respect than an aural one, as with the former the personal element does not enter into the calculation to any extent, nor is the ear capable of giving accurate quantitative measurements.

The average value of a recurring wave shape whose half-cycles match each other is zero, that is, the mathematical addition of the positive and negative halves of the cycles gives zero as their sum. The average value of any recurring wave shape whose half-cycles do not match each other in amplitude must be greater or less than zero.

A direct-current ammeter of the usual type reads only average values of current. Suppose such a meter is inserted in the plate circuit of a tube. If a signal is introduced to the grid of the tube and the amplified output is free from harmonic distortion, the average value of the plate-current change produced by the incoming signal will be zero. Thus the average value of the plate current is the same whether the signal is present or not, and, as the meter in the circuit reads average values, its indication will remain unchanged.

If the tube is not operating linearly and is introducing harmonics, the average value of the audio-frequency component in the plate current will be greater or less than zero. It will then add or subtract from the meter reading obtained when there is no signal present.

If the signal is not steady but varies in level as does a program, the meter reading will either dip or rise at the high-level peaks.

Such a direct-current meter is usually kept in critical circuits in order that the presence of distortion may be immediately indicated. The remedy is, of course, to reduce the signal input.

216. Shot Effect.—With the advent of the high-gain amplifier, capable of reproducing frequencies up to 15,000 cycles, the so-called *shot effect* in tubes becomes very noticeable. This is a steady high-frequency hiss in the tube output, owing to the fact that the instantaneous emission

of electrons from the cathode is not perfectly regular but spasmodic in character. The irregularities in emission occur at a high audio frequency and thus may be heard in any loud-speaker capable of reproducing frequencies of this order. Some tubes have this defect to a greater extent than others, even among the same type. For instance, in a certain amplifier designed to use type-37 tubes, it was found that only one tube out of three, on the average, was suitable for use in the input stage. With some tubes the effect may be somewhat regulated by control of the heater or filament current.

The shot effect is not ordinarily apparent when using low-gain equipment, as the signal input to the grid of the first tube is relatively so large that the hiss in the plate circuit is kept well below program level, and, during periods in which there is no signal input, the amplification is not great enough to make it noticeable.

In high-gain amplifiers the signal input is very minute and the program level may not be relatively high enough to mask the shot effect. Also the amplification of the hiss from the first tube will be much greater, rendering it very apparent during silent periods in the program. Tubes have been developed which minimize this condition and render them suitable for use with high-gain equipment.

217. Advantages of Low-gain Amplifiers.—As the number of stages in an amplifier is increased, it becomes more and more difficult to secure stable operation. Owing to undesired energy being fed back from output to input, through coupling of the transformer fields or otherwise, the circuit has a tendency to go into oscillation. The precautions necessary to prevent such action become prohibitively expensive and awkward when more than two successive steps of high-gain amplification are employed. With a low gain per stage, stable amplifiers having three or four stages may be built. For this reason, when it is necessary to construct reproducing systems having great overall amplification, instead of concentrating the entire gain in one amplifier, it is split up over several succeeding units of individually low-gain amplifiers of high stability. These are placed sufficiently far from each other to avoid undesirable coupling between them.

This system has the added advantage that attenuation networks or gain controls can be inserted between the units to cut down the overall amplification. These, by increasing the loss between the individual amplifiers, decrease the possibility of sufficient feed back between them to start the system oscillating.

If the individual units are run from the same power supply, it is necessary to provide adequate filtering in each amplifier to prevent audio-frequency variations from entering into the supply circuit and so affecting the stability of the system, or causing cross talk in other units energized by the same power source.

There are three types of amplifiers used in broadcasting, namely, the resistance-, impedance-, and transformer-coupled types. They will now be discussed primarily as they apply to studio amplifiers.

218. The Resistance-coupled Amplifier.—This type of amplifier (see also Sec. 139 *a* and *d*) is used when high voltage is available, excellent frequency response a necessity, and space at a minimum. Thus it is the ideal amplifier for use with the condenser microphone. Such a circuit is shown in Fig. 216.

In order to secure the maximum percentage of amplification from the tube, the value of the coupling resistor R_6 in the plate circuit is kept as high as possible. Its maximum value is, of course, at once fixed by the plate current of the tube and the voltage drop that can be permitted across it. For example, in the circuit given, the plate current of the tube is 1 ma. with 90 volts on the plate, and, as the maximum voltage available is 200 volts, there remains an allowable drop of 110 volts across the resistor R_6 . As $R_6 = E/I_p = 110,000$ ohms, this is the maximum value of the coupling resistor.

As previously stated, in order to keep audio-frequency currents out of the series-filament circuit and plate supply, the tube should be either series- or parallel-fed. There is shown in this circuit diagram an application of series feed using a filter resistance R_6 instead of a filter choke. By not using a choke, space is conserved, the picking up of stray fields in a low-level circuit is avoided, and expense is reduced.

By sacrificing a very small part of the possible amplification (realized μE_g) of the tube we may take 10,000 ohms from the optimum value of the coupling resistor R_6 and use it in conjunction with a 1-mf. condenser as a filter circuit, C_3, R_6 , as shown in Fig. 216. Examining the efficiency of the device we find that, if the amplifier is to respond to frequencies from 20 to 10,000 cycles, at 20 cycles the condenser must have a reactance of about 8,000 ohms, and thus only about 55.5 per cent of the 20-cycle current will be filtered out of the desired circuit. At 40 cycles the condenser must have a reactance of 4,000 ohms, and the efficiency will rise to about 71.5 per cent. At 80 cycles the condenser reactance must be about 2,000 ohms and the efficiency will then be about 83 per cent. As the frequency increases, the efficiency rises, so that still better care is taken of the higher frequencies. Such a filter is quite good enough to serve the purpose required.

An examination of the circuit will show that the coupling condenser C_4 and the grid resistor R_7 form a voltage-dividing circuit across the plate resistor. This determines what part of the audio-frequency voltage developed across that unit shall be passed on to the grid of the next tube.

In order to realize the greatest possible gain, the combined impedance of the condenser and grid resistor must be kept so high that the total load resistance in the plate circuit is not materially lessened by their shunting effect on the coupling resistor. As only the voltage appearing across the grid resistor is actuating the next tube, the reactance of the condenser must be kept low in comparison with the grid resistor, at every frequency, if maximum amplification and good frequency response are to be obtained. Carefully designed and constructed amplifiers of this type have been built for use in television circuits and have produced a flat frequency characteristic from 15 to 45,000 cycles. The higher the frequency response desired from such an amplifier, however, the lower the gain it is possible to obtain from it. As the frequencies entailed increase, the reactances of the inter-electrode capacities of the tube and the stray-wiring capacities in the circuit decrease to such a point that these capacities which are in shunt with the grid resistor act to lower materially the impedance of

the grid circuit. To minimize this effect it is necessary to lower the value of the grid resistor, as the desired frequency response is raised. This lowers the effective load in the plate circuit with a consequent decrease in amplification.

With resistance coupling it is of course impossible to ever realize the full amplification power of the tube, consequently this system offers a lower gain per stage than any other.

219. The Impedance-coupled Amplifier.—This type is able to work with lower values of plate voltage (see Sec. 139*e*) than the resistance-coupled amplifier. As supplies of from 350 to 400 volts are well standardized and are required for the power tubes operating monitor loud-speakers, this point is not so important in the control room as it is in the field. This type of amplifier is shown in Fig. 149(*a*) and (*b*).

To insure good high-frequency response, the distributed capacity of the coil must be kept low, because it acts as a shunt reactance across the coil and decreases as the frequency increases and vice versa.

To insure good low-frequency response the inductance of the coil must be kept high so that, even at the lowest frequency to be reproduced, its reactance will be many times higher than the plate resistance of the tube. It sometimes proves difficult to secure a large enough inductance to insure the desired low-frequency response. When this happens, advantage can be taken of the fact that a series-feed filter condenser, together with the coil, forms a series-resonant circuit (see L_p and C_1 , Fig. 237(*a*) or (*b*)).

By choosing a condenser of such value that it resonates with the coil at the frequency where the response has fallen, the total impedance of the circuit will become lower as the frequencies approach this point. Then the audio-frequency current through the circuit will increase correspondingly, causing a greater audio-frequency voltage drop across the reactance of the coil than could otherwise occur.

If the possibility of increasing the low-frequency response of the amplifier by the method just described is not taken advantage of, care must be exercised to choose a condenser of such value that it does not resonate with the coil at some undesired frequency. When replacing a defective condenser in this position in an amplifier, the operator should remember this and insert a unit of the same value. Generally, one of larger capacity should be used if possible.

The impedance-coupled amplifier has a low gain per stage, as no amplification other than that in the tube is developed. (No transformer step-up is obtained.)

In equipment having high overall gain it is highly desirable to avoid noisy tubes. Unfortunately for the purpose of the impedance- and resistance-coupled amplifiers, the high- μ tube is inherently microphonic, this defect severely limiting its use in broadcast amplifiers. Thus, when used in reproducing systems, the gain it is possible to realize with such amplifiers is still more limited by the necessity of having to use non-microphonic low- μ tubes to handle the extremely small inputs afforded by modern microphones.

Impedance-coupled amplifiers having an overall gain of 70 db and employing a high- μ tube in the input stage have been used successfully to handle the output of carbon microphones.

220. The Transformer-coupled Amplifier.—Owing to the great improvement which has taken place in the characteristics of audio transformers, more recently designed amplifiers make considerable use of transformer coupling. The great advantage of this system is the high gain per stage it is possible to obtain by its use (see Sec. 139*c*).

Just as with the coupling coil of the impedance-coupled amplifier, the primary-inductance value of the audio transformer must be kept high in order that the voltage

across it at the lowest frequency encountered will not differ from that encountered at the highest, to any marked degree.

The step-up given to the signal by the transformer depends upon its turns ratio, and it is desirable to keep this as high as possible. Unfortunately, the greater the number of turns on the secondary winding, the greater its distributed capacity, with a consequent falling off in the high-frequency response obtainable. There is, therefore, a definite limit to the number of secondary turns that can be used, and still have the transformer possess a good high-frequency response. As it is necessary to have a large primary inductance (on the order of 300 henrys), this calls for a great number of turns on this coil and so restricts the allowable turns ratio to a low value. If, in any way, the necessary inductance of the primary could be achieved with fewer turns, the transformer ratio would immediately benefit. It is for this reason that all modern high-quality transformers use high-permeability cores. Such alloys are usually a composition of nickel and iron which have been given special treatments in their formation. By greatly increasing the number of electromagnetic lines of force, through the core material, for the same impressed magnetizing force, they allow the use of a much smaller primary coil while maintaining the same inductance.

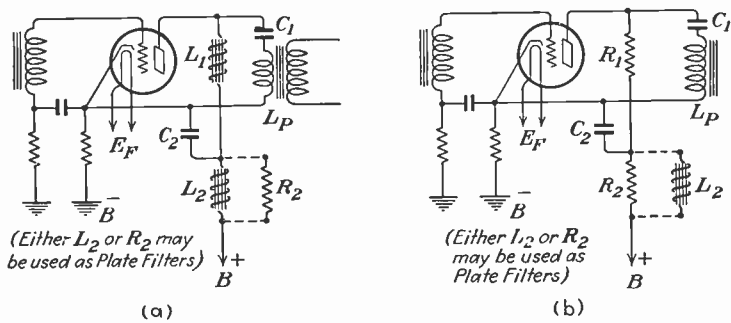


FIG. 237.—Methods of parallel-feeding in audio-frequency amplifiers.

The higher the permeability of the core material, and the greater the inductance of the winding, the greater the ease with which the core will become magnetically saturated by the passage of small direct currents through the windings. In order to maintain maximum primary inductance, it is necessary to keep the direct current in the plate circuit out of transformer coils.

This is best accomplished by the use of parallel feed. A suitable circuit is shown in Fig. 237(a). The direct current flows only through the coils L_1 and L_2 . The inductance L_1 has a reactance which is high compared to the impedance of the circuit comprised by the condenser C_1 and transformer primary, at every frequency involved. The audio-frequency current then has two paths over which it returns to the cathode. As that which includes the transformer primary offers the least resistance, the main body of alternating current will travel through it. The coil L_2 and the condenser C_2 prevent the audio frequency which is returning through L_1 from passing through the power-supply circuit. Care must be taken to prevent the coil L_1 and condenser C_2 from offering low impedance at some undesired frequency owing to resonance. C_2 is therefore chosen as of a high enough value to resonate with L_1 at a frequency well below those being reproduced. L_2 also serves as a filter for any irregularities in the power supply to the plate circuit such as the alternating-current ripple present when using a rectified current or the direct-current ripple present if the supply is obtained from generators.

Another parallel-feed circuit is shown in Fig. 237(b). If the tube has low plate-current and voltage requirements, and there is an excess of voltage available at the

supply source, advantage may be taken of the circumstances to replace the coil L_1 in (a) with a resistor R_1 as in (b). The value of such a resistor is given by the allowable voltage drop divided by the plate current. The value of this resistor must be high in order to secure efficiency.

In all parallel-feed circuits the value of C_1 is chosen to resonate with the inductance of the transformer at the lower frequencies where the transformer-response curve falls owing to its reactance becoming comparable to, or lower than, the alternating-current plate resistance of the tube. By this means larger currents appear in the coil than would otherwise be possible, as detailed in Sec. 219 on Impedance Coupling. The resonant peak is prevented from becoming too sharp by the ohmic resistance of the transformer winding. The response of the system is prevented from becoming pronounced at any one note by choosing a different low frequency at which to resonate each stage in the system. In the diagram (a), as the frequency is lowered, the reactance of coil L_2 is also lowered, and its efficiency in causing most of the audio frequency to flow through the transformer is correspondingly decreased. The value of the resistance replacing it in the second circuit (b), however, is independent of frequency. Hence, by its use, considerably better low-frequency response may be obtained, as it is more efficient in shunting the current through the transformer primary at low frequencies than is the coil. At high frequencies it also has the advantage of stabilizing the circuit by its constant value and lack of resonating effects with stray circuit capacities. Its use assures a more uniform frequency response at less expense than is possible with the coil. It also avoids the possibility of picking up stray fields through the coil inductance.

If the supply circuit is sufficiently free from disturbances, and the permissible value of the resistance R_1 is high enough, part of it may be sacrificed to become part of the filter resistance R_2 .

Parasitics.—Due to the distributed capacity and inductance present in the transformers, systems employing them sometimes have a tendency to go into oscillation at high frequencies, often above audibility. If the operation cannot be rendered stable by proper filter circuits it is sometimes necessary to insert high values of resistance in series with the grid circuits to prevent these *parasitic oscillations*. Another means of prevention is to load the transformer secondary with a high resistance, but this tends to decrease the high-frequency response. Push-pull circuits are especially liable to parasitic oscillation. This condition may be prevented by the insertion of a small inductance or, if the voltage drop is permissible, a resistance in the plate circuit immediately following the center connection of the plate transformer.

Owing to their greater gain per stage, more care must be exercised to insure stability in transformer-coupled amplifiers than with any other type. More than two stages per unit are rarely used. As an illustration of the gain made possible, preamplifiers for dynamic microphones have been built which have a gain of 40 db with only one stage of amplification. As an illustration of the excellent frequency response obtainable, commercially used amplifiers have been built which use five stages of transformers and are essentially flat in their response from 20 to 15,000 cycles. There are transformers designed for use in television circuits which are capable of transferring frequencies up to 40,000 cycles per second.

221. The Main Gain Control.—With any system of amplification the placement of the main gain control is an important point. If high-level mixing is used, the main amplifier seldom has a gain of over 70 db, and the gain control may be placed directly at the output of the mixer without encountering undue difficulty in maintaining quiet operation. If low-level mixing is used, the main gain control is usually placed between the low- and high-level amplifiers. This position has two advantages. First, because the unit is working at comparatively high program levels, it will be quiet in operation; and second, because of its position, it decreases the overall amplification between input and output of the amplifier system and thus vastly increases the stability.

In order not to disturb the impedance relations in the circuit, such gain controls are made up of variable, constant-impedance attenuation networks, on the order of those illustrated in Figs. 232(c) and (d) and 234. They must be non-inductively wound so that they will not be affected by stray fields. Their capacity and unbalance to ground should be kept

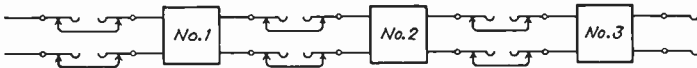


FIG. 238.—Method of connecting units of studio amplifiers together through jacks.

low, so that the frequency response of the system is unaffected by their use.

If the gain control employs a detent and is mounted on the same rack as the first amplifier, trouble may be encountered through tube microphonics occasioned when the control is moved. This may be helped by removing the detent, but a far more convenient method is to mount all switches and controls on a turret apart from the amplifier. This turret, which also carries the volume indicator, is mounted on a desk so placed that the operator is looking into the studio through the soundproof plate-glass windows provided for this purpose.

The use of potentiometers to act as volume controls, by varying the voltage input from the secondary of the transformer to the grid of the tube, is sometimes encountered. This system is becoming obsolete as the control cannot be made remote to the amplifier owing to the high impedance of the leads and because it has been found detrimental to good high-frequency response.

222. Bridging Circuits.—To insure maximum flexibility, and to enable service to be resumed with minimum delay in case of breakdown in any unit, all the inputs and outputs of the component parts of the reproducing system are brought out to jacks. They are normally connected together at the jack contacts as illustrated in Fig. 238.

Thus in the circuit shown, if unit 2 should prove defective, it may be bridged out of circuit by inserting a patch cord into the output jack of unit 1 and into the input jack of unit 3. The action of the plug inserted

into the jack is to lift automatically the jack arm and break the contact with unit 2, removing it from the circuit entirely. In a like manner, duplicate pieces of equipment may be brought out to jacks, and if trouble occurs, all that is necessary to substitute workable units into the circuit and to remove the defective part is a pair of patch cords.

As all equipment is in this way brought out to a central panel, long interconnecting leads are necessary. In order to minimize the effect of their distributed capacity upon the frequency response, the input and output impedance of all units is kept at a low value. This also helps to prevent undue losses to ground through stray capacities and stray pick-ups due to the fields about the cables and patch cords. Usual terminal impedances are 50, 200, 250, and 500 ohms.

In order to eliminate cross talk and pick-up by the circuit, all exposed wiring is shielded. It is of importance first to band together all the individual conductor shields and then to ground these as a unit at as many points as is convenient. If this is not done, any differences in potentials to ground of adjoining shields are apt to cause trouble when using the extremely high gain encountered with modern equipment.

223. Power Supplies. *a. High-voltage Supply.*—In some installations the high-voltage storage batteries used for plate supplies have been replaced by alternating-current rectifiers. Earlier types utilized 50-watt amplifier tubes as rectifiers with plates and grids tied together. They are used in a conventional full-wave circuit followed by a multi-section filter of ample proportions to insure a quiet output. Later equipment makes use of the hot-cathode mercury-vapor rectifier tube, which is ideally suited for the purpose. In such circuits the rectifier output is worked into an inductive load, as, with a capacitive input to the filter, there is a tendency for unbalance to occur and for one tube to take the full load.

In these installations such ample provisions are made in regard to filter requirements and adequate current supply that one tube, working as a half-wave rectifier, is capable of supplying the full current for the system, without ripple appearing in the filter output. In this way interruptions to service owing to a faulty rectifier tube is minimized.

In order to avoid voltage fluctuations when using these systems it is generally necessary to leave the full load on at all times or to provide means for changing the rectifier output, as the load changes. A bleeder resistance is always connected across the filter input or output to prevent the voltage rising to such abnormal values that it might injure the condensers if through accident the load is removed while the unit is in operation. Gaseous-type tubes sometimes generate high radio frequencies which prove objectionable if allowed to enter the high-voltage power-supply system. To prevent this, a low value of condenser, usually about 0.1 mf. or less, is connected between the plate and cathode

of such tubes, in order to act as a short circuit for the radio frequencies.

When mercury-vapor tubes are first placed into service, or replaced after being removed from their sockets, they must be warmed by lighting the filament for 15 min. before applying the plate voltage. This distributes, in the form of vapor, any spots of mercury which might be adhering to the filament. After this initial warming, it is only necessary to light the filament for from 15 to 30 sec. before applying the plate voltage each time the rectifier is placed into service. Provision is made to do this either manually or by means of time-delay relays in the rectifier equipment.

b. Filament Batteries.—When batteries are employed as a source of filament-current supply, precautions must be taken to prevent voltage fluctuations under varying loads. By using batteries capable of handling peak demands and by the installation of conductors of ample size, in order to assure such a small ohmic resistance that a negligible voltage drop occurs through them under full load, much can be done in this respect.

In the better installations the batteries are maintained in a fully charged condition at all times. This is accomplished by floating them across a direct-current generator, as explained in Sec. 399 on Charging Methods, which is capable of supplying the full load for the circuit. In this way an absolutely steady voltage output is maintained because the moment the batteries tend to fall to a lower potential than the generator voltage, they automatically begin charging. The generator is normally carrying the full load of the circuit. If through an emergency condition the generator should cease functioning, it is automatically cut out of circuit and the storage batteries assume the load without any interruption to service taking place.

The irregularities in the output of the direct-current generator, due to commutation, must of course be filtered out. As the current in the circuit is large, high values of inductance cannot be placed in series with it, since the cost of making reactors, with cores sufficiently large to prevent saturation and conductors heavy enough to prevent excessive voltage drop across them, becomes prohibitive. With such small reactances available for filtering purposes, it would be necessary to use extremely large capacities in conjunction with them in order to secure adequate filtering. Fortunately there is a more effective and simpler way of accomplishing the required result.

The battery which is floated across the generator takes the place of the conventional capacitor and, owing to its extremely low internal resistance, is much more effective in its action than the former. In this way any irregularities in the current supply are completely filtered out. This same principle of using a battery in a filtering circuit appears in many instances, and it is worthy of note that a battery can and does pick up stray fields, just as does a coil, and it is therefore sometimes necessary to move the battery around in various positions to lessen this effect.

In less elaborate installations a manually operated variable resistance is used in series with the main line. As the battery voltage falls in service, adjustment of the rheostat must be made in order to keep the line voltage constant in value. Because of the comparatively high resistance of the supply line, the load on the circuit is maintained as steady as possible. During the period the equipment is not in service the batteries are placed on charge.

224. Biasing Voltage.—Growing out of telephone-plant practice, the usual source of filament current for battery-operated amplifiers is a

12-volt bank of storage batteries. In order to utilize this power economically, most broadcast equipment is designed to operate with series filaments. In the earlier stages advantage is taken of the voltage drop in the filament circuit to furnish bias for the tubes. The tubes which must be supplied with higher negative grid potentials than are available are biased by means of small dry cells which are cased as an integral part of the amplifier. If heater-type tubes are used, advantage of this is taken to bias them by means of the voltage drop across a suitable cathode resistor placed in the plate-return circuit (see Sec. 90).

In Fig. 239 is illustrated a circuit obtaining its bias by means of voltage drops in the filament circuit. The bias on the grid of the first tube is the difference in potential between the grid end of the resistor R_1 and the negative end of the filament. Similarly the drop across R_1 , R_2 and the first filament F_1 furnish bias for the second tube.

The bias for the last tube shown is made up of the combined voltage drops across the resistors R_1 , R_2 , and R_3 , and also those across the filaments of the first two tubes. It will be noted at once that any variation in filament current will immediately change the bias on all tubes and so seriously affect the operation of the amplifier.

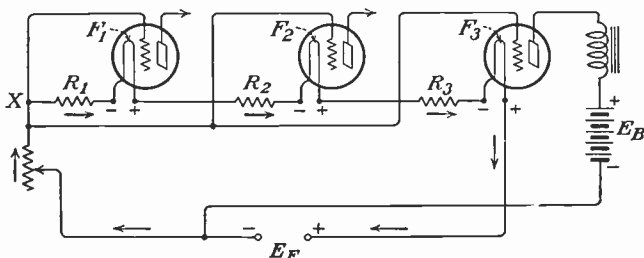


FIG. 239.—Method of obtaining bias through filaments in series and resistors.

In such amplifiers the filament rheostat shown must be capable of varying the current with the utmost smoothness, for any sudden variation or any change occurring at an audio frequency will impress itself upon the grids of the tubes as voltage variations and so be amplified by them. It is, of course, impossible to obtain a greater bias by this means than the impressed voltage from point X less the filament voltage of the tube at the positive end of the series. Therefore, the battery E_F must supply voltage at a value equal to the sum of all the bias and filament voltages combined. Let us trace the plate current for the last or power tube. The electrons leave the filament, reach the plate, pass through the high-voltage plate supply, thence through the series filaments F_1 , F_2 and resistors R_1 , R_2 , and R_3 to F_3 thus completing the circuit. In the output tube the plate-current variations may be a considerable fraction of the filament current. This plate current, varying at the program frequencies, is passing through the resistors which are furnishing bias to the amplifier tubes. Their bias then, instead of being steady, is rising and falling at program frequencies and causing a consequent rise and fall in the plate current of the tubes. Energy is being fed back into the circuit from its output stage. Whether this will cause regeneration or degeneration depends upon the number of tubes and the circuit constants; in any case, its effects will be highly undesirable. If the plate currents of the remaining tubes are also traced, it will be found that they too exert this effect but to a lesser degree.

The plate-current variations must be removed from the filament circuit to insure proper operation of the amplifier. This is done by either a *series* or a *parallel* feeding of the individual plate circuits. An example of series feeding is shown in Fig. 240. The audio-frequency variation, after passing through the primary of the coupling transformer, or through the plate resistance if it is resistance-coupled, or through the plate reactor if it is impedance-coupled, has before it two paths enabling its return to the filament whence it originated. The filter choke coil L presents a very high reactance to it and the condenser C a reactance which is by comparison low. Practically all the audio-frequency current will return to its filament via the condenser, and the slight amount passing through the choke and thence through the biasing resistors will be reduced to harmless proportions. It must be noted that to make such circuits really effective, the by-passing condenser should be returned directly to the filament which the current through it is trying to reach. Grounding the return terminal of such a condenser or connecting it to the negative high-voltage lead is bad practice and defeats the purpose of the circuit. It should be noted that such filtering also keeps the audio frequency out of the plate-supply circuit. Parallel feed is taken up in detail at a later point. It may serve the same objective but with added advantages.

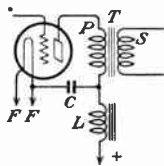


FIG. 240.—
Method of connection for series-feeding of plate circuit.

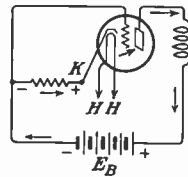


FIG. 241.—Method of obtaining grid bias with cathode-heater tubes.

The second system mentioned for obtaining bias is shown in Fig. 241. Tracing the plate current, we find the electrons flowing from cathode to plate, thence through the high-voltage supply and through the resistor back to the cathode. The consequent voltage drop across the resistance makes the grid negative with respect to the cathode. The calculation for the proper value of resistance is very simple:

$$R_{bias} = \frac{\text{desired bias volts}}{\text{d.-c. plate current}}$$

Of course, the impressed high voltage must be equal to the sum of the plate voltage and the drop across the biasing resistor.

Let us find the effect of alternating-current plate-current variations in such a circuit. Suppose the grid to be swinging toward the positive with the incoming signal; the plate current through the resistor will rise, increasing the voltage drop across it and hence biasing the tube more negatively. As the current effect travels through the circuit with nearly the velocity of light, the effect is practically instantaneous. Hence the effective values of the positive cycle of the incoming signal is decreased by the increase in negative bias which it engenders. The negative half of the cycle will cause a decrease in plate current and hence a drop in the biasing voltage and a consequent partial nullification of the signal.

Thus the presence of the signal variations in the biasing resistor causes a degenerative effect and a loss in amplification. The situation may be relieved in patchwork fashion by placing a suitable value of condenser across the biasing resistor; this will offer a lower impedance path to the signal and hence a lessening of the audio-frequency

voltage drop across the combination. To be effective the value of the condenser must be very large, and, as its reactance changes with the frequency, it will affect the frequency response of the circuit. The proper solution to the problem is to employ plate filtering (Fig. 240) in the plate circuit, and thus eliminate the variations through the biasing resistor.

225. Master Control Room.—In large installations all amplifiers and their accessory equipment are kept in a central master control room, only the turret housing the mixer equipment, main gain control, and volume indicator for each individual studio appears in the subcontrol room, adjacent to the studio. By this practice, when using batteries, the power wiring is considerably simplified.

IV. ALTERNATING-CURRENT OPERATED EQUIPMENT

226. Advantages of Alternating-current System.—The greatest step forward in program-transmission systems is the advent of equipment capable of being operated directly from the 110-volt 60-cycle alternating-current mains. With such equipment each subcontrol room and its adjacent studio are usually made complete, self-contained and independent units, comprising complete transmission and monitoring systems in themselves. The elaborate and costly wiring necessary with a centrally located battery supply is done away with, as are also the maintenance expense and the additional floor space which the use of batteries entails.

When batteries are employed, it is sometimes necessary that all equipment be left with the power supply on, whether in use or not, to prevent voltage fluctuations from occurring and affecting other equipment which is being used. This handicap is done away with when operating on alternating current, as there is incorporated in each unit an extremely efficient voltage regulator capable of keeping the voltage constant under widely varying power-line voltages. Such instantly effective and efficient regulators cannot be made for use with direct-current-operated equipment.

With master control-room systems all amplifiers are kept in one master control room, only the mixer and its associate equipment appearing in the studio control room. The low levels involved in transferring the mixer output to the master control room necessitate that especially quiet individually shielded cables be used. This prevents cross talk from other adjacent cables carrying higher level programs destined either for audition, transmission line, or the control-room-monitor speaker. When alternating-current-operated equipment is used the program leaves the individual control room at a level of 0 db ready to be impressed directly upon the transmission line. In this way much of the otherwise necessary wiring is eliminated. This eliminates, or greatly reduces, the functions of the master control room.

The equipment for the studio control room is a complete transmission system in itself including the power supply, assembled as a unit upon a rack or racks. This makes the apparatus highly portable and easy and inexpensive to install. It assures ease of maintenance and quick location of any troubles that might arise, as all necessary equipment is centrally located and within reach of one person.

227. Amplifier-tube Developments.—Tubes having their heaters operated by alternating current have long been familiar in receiving equipment. Even the best of such tubes, however, cannot be used in high-gain, low-input amplifier equipment without introducing serious hum into the circuit, due to the action of the tube itself. Thus, before the development of a special tube, which will be described later, it was necessary to rectify the alternating current before applying it to the heaters when using such tubes in studio equipment.

Indirectly heated cathode-type tubes are always used because they offer individual and isolated plate-circuit returns, have a tendency to be non-microphonic, and, most important of all, allow the heater supply to contain a large amount of ripple without introducing disturbances in the tube output. This allows a minimum of filtering to be used in the supply circuit. The rectifier for supplying direct current to the heaters is usually a gaseous-type tube capable of handling large currents at low voltages. It is used in a conventional full-wave circuit followed by a filter of sufficient efficiency to suit the purpose at hand (see Sec. 321).

There has been developed an indirectly heated cathode-type vacuum tube capable of being operated in low-gain circuits while being heated directly by alternating current. The noises in the output circuit of this tube are more than 20 db below those of the most quiet alternating-current-type receiving tubes. They also have the advantage of being exceptionally non-microphonic, being in this respect some 10 db better than the best non-microphonic filament-emitter-type tube. Their use, of course, does away with the necessity for the filament-current rectifier and considerably lessens the cost and bulk of the equipment.

228. Reducing Hum Level.—As the fields about the wires carrying the alternating-current filament supply may induce hum into the system, every precaution must be taken to keep its value low and to keep it as far as possible from the equipment. If the two conductors carrying the current could occupy the same point in space at the same time the fields about the conductors would completely neutralize each other as the currents in the wires are equal and flow in opposite directions. This hypothesis is manifestly impossible, but it is approximated by running the wires as closely together as possible, that is, by twisting them. As the fields are not completely neutralized the alternating-current wiring is enclosed by a heavy shield, and is carried on the opposite side of the rack from the program circuits.

The filament transformers and high-voltage rectifier are mounted at the bottom of the rack and a heavy magnetic shield interposed between them and the rest of the apparatus. The student must remember that, contrary to the general impression, a perfect shield is possible only in theory. It is for this reason that the low-level input amplifier is kept at the extreme top of the assembly, away from the fields of the power-transformer.

Advantage is taken of the high voltage available to utilize resistors in the individual plate circuit of each tube as filters and for parallel-feeding the transformers. The condensers following the resistors in the circuit have comparatively low voltages across them when plate current is flowing.

229. Time-delay Relay.—When the equipment is first turned on, however, it takes about 30 sec. or more for the cathodes to become hot enough to allow plate current to flow. During this time there is no voltage drop across the series resistor, and consequently the full rectifier voltage appears across the low-voltage coupling condensers. To prevent this and also to allow the filaments of the mercury-vapor high-voltage-rectifier tubes to become warm, means are provided to energize the high-voltage transformers separately, after the filament circuits have been closed a sufficient length of time. This is usually accomplished automatically by means of a time-delay relay.

When the power switch is closed, it energizes the filament circuits and also causes current to pass through a coil of resistance wire wrapped about a metallic arm, but insulated from it, which carries a contact point at its end. The heat created by the current through the resistance coil causes the metallic arm to expand in such a way that its end describes a partial arc and closes a contact. The length of time this takes can be nicely regulated by the current applied and by adjustment of the space between the contacts. This closed contact completes a circuit connecting an alternating-current relay to the line, which upon operating closes the primary circuit of the rectifier plate-supply transformer. It locks itself up in this closed position by shorting the thermal relay contacts, until such time as the power switch is opened when the circuit through the heating resistor is opened releasing and opening the relay.

230. Voltage Regulation.—If a rectifier similar to that described above is used, it is not supplied with power directly from the line but through a voltage-regulating transformer the action of which is instantaneous and efficient and which is described in Chap. XIII on Rectifier Units. Such regulators are very carefully designed for the load that is to appear across their outputs, hence this load must not be changed by the addition or subtraction of equipment, nor should tubes having a heavy current drain be removed while the power supply is carrying the load.

It is important, if the equipment is being operated by means of rotary converters, to insure the maintenance of proper supply frequency, otherwise the voltage-regulating device may act as a step-up or step-down transformer. When properly designed, such regulators are capable of supplying constant output voltage over varying inputs ranging from 90 to 135 volts. Their use safeguards the equipment, greatly reducing the wiring costs in large installations, and increases the flexibility of the unit.

231. Amplifiers in Alternating-current-operated Equipment.—The amplifiers used in these equipments are in every respect similar to those already taken up, with the exception of the added precautions taken to

prevent stray fields and ripple from the power supply from entering the circuit. Ripple-filtering resistors are sometimes replaced by high-inductance coils, to increase the efficiency of the circuit. It is most important to prevent the alternating-current ripple in the plate circuit from appearing on the grid through the agency of the biasing resistor. To accomplish this an ingenious filter circuit is used which also serves to prevent any degenerative effects owing to audio frequency through the bias resistor.

In the output stages where *power* must be supplied, heater tubes are relatively inefficient. It is customary to use, for power purposes, tubes employing oxide-coated filaments operated on alternating current. These tubes are always used in a push-pull circuit which tends to balance out their hum. The input levels are kept so high that the hum is always so far below program level that it is negligible. Thus, even if the full output power of the combination is not needed, the signal input to the grid is always kept at a maximum and the output stepped down to the desired value by attenuation networks. This has the further advantage of reducing the output hum of the power stage by the value of the pad following it if the program is interrupted.

V. PROGRAM CONTROL AND TRANSMISSION

232. Monitor Amplifiers.—There are several systems of monitoring. One of these systems is to make the line amplifier capable of supplying a level of approximately 25 db or more, and placing a loud-speaker across its output; then, by means of a pad, which is also across the output, to step this value down to line level. This system has a great disadvantage in that the power available for the speaker is fixed and cannot be readily changed. Room noises affect the volume produced by the monitor loud-speaker to a very great extent; thus, in a quiet room the loud-speaker may give a satisfactory volume, whereas, if several people are present and causing room noise, the sound level may be annoyingly low. Also if the operator suspects some unwarranted noise in the output he should be able to increase the speaker volume at will, in order to make possible its detection.

It is for this reason that separate amplifiers are desirable for monitoring purposes. As the amplifier input is to bridge the line it has a high impedance, from 4,000 to 20,000 ohms, in order that it will not draw an appreciable power from the main load.

The gain and output power depend upon the number of speakers being fed and the method of feeding employed. If the amplifier is to supply all speakers directly, it should be capable of supplying each unit with a power level of about 25 db. If more than one or two speakers are used with this system, a very large output is required. If they are scattered about the building, very high power levels must be impressed

upon the cables connecting them to the amplifier. This makes the prevention of cross talk in neighboring circuits a difficult matter. In order to provide for changes to be made in the individual speaker volume, without affecting others connected to the same supply circuits, a constant-impedance attenuation network must be used between each speaker and the line. Such a network is shown in Fig. 234.

The larger the number of speakers supplied, the greater the problem of cross talk. Under these conditions it is much better to use a line-bridging monitor amplifier the output of which supplies a moderate power, such as 0 db, to a low-impedance bus which runs to all the speakers. Each speaker is then equipped with an individual amplifier, bridging the bus and capable of raising the volume to any desired level. The manipulation of the volume control which is incorporated in a speaker amplifier will not affect the program level at any other speaker. The use of the line-bridging amplifier feeding the bus serves to isolate the speaker circuits from the line and prevent any short or open circuit in them from affecting it. This system also provides much greater flexibility as speakers may be added or removed from the circuit, as conditions may require, without noticeable effect on the system.

Line-bridging amplifiers have high-input impedances and are designed to be energized by zero level and give an output of from +10 to 20 db. This output is stepped down to line level by a suitable pad between amplifier and line. The advantages of using a pad in this way are discussed under Attenuation Networks (Sec. 238).

233. Volume Indicators.—In order to allow the operator to control properly the gain of the system, he must be aware of the exact program level at any instant. For this purpose, a meter, calibrated in decibels and connected across the output of the amplifier, is mounted near the main gain control.

Such a meter must read the peak values of the audio-frequency wave and must be capable of very quick response to enable it to follow the rapid variations in program level. If the audio-frequency wave is first rectified and then passed through a suitable direct-current meter, these conditions will be fulfilled.

a. Vacuum-tube Voltmeter Type.—In Fig. 242 is shown one method of doing this. The tube is biased almost to the point of cut-off, hence it will amplify only the positive half-cycle of the signal on the grid and acts as a rectifier. By means of the potentiometer R , the bias is adjusted so that there is enough direct-current plate power flowing through the meter to make it read about one-twelfth of the full scale. The purpose of this is to prevent the needle striking the case below its non-operating

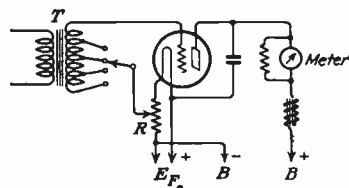


FIG. 242.—Vacuum-tube voltmeter.

position due to its impetus on the return from the upswing caused by a high-level program peak.

The current through the meter is the rectified audio frequency and, as the needle inertia integrates the more rapid audio-frequency variations, the indicator pointer will follow the more gradual rise and fall in program volume.

The input transformer *T* is tapped so that the device may be made to read different input levels. The usual method of calibration is to have the meter marked 0 db somewhere about the center of the scale, the points immediately above and below it being marked with positive and negative values, respectively. The position of the arm making connections to the transformer taps is also marked in decibels. The actual value in decibels of the program level at any instant is given by the sum of the arm and meter readings. For instance, if the arm is set at 0 db and the meter reads +2 db, the program level is +2 db. Or if the arm is set at +10 db and the meter reading is -2 db, the program level is +8 db.

The circuit shown in Fig. 242 is a general one, modifications in it being often made. The filament may be alternating-current operated or the input voltage may be regulated by means of a potentiometer across the transformer. In any case the principles involved will be the same. The meter is usually a 1 or 1½ ma. direct-current meter.

b. Copper Oxide Rectifier Type.—The second method of rectifying the audio frequency before impressing it on the meter has the advantages of great simplicity, small space requirements, requiring no power supply, and comparatively low cost. It is based on the use of a recently developed type of copper oxide rectifier the output of which is independent of the impressed frequency up to values higher than is necessary for the purpose at hand. Such units are used in a bridge circuit to rectify the current for a direct-current meter which is similar to that described. As they are quite small, the elements are housed in the meter (usually a 3-in. scale size), and it is thus made into a complete unit. Thus there is available a peak-reading milliammeter, which by means of suitable multipliers can be converted into a peak-reading voltmeter, with a resistance of about 5,000 ohms per volt. The meter then has a high enough impedance to be bridged across the line without noticeably affecting the program level.

By arranging a switching circuit which changes the value of the multipliers in series with the meter while at the same time keeping its impedance to the line the same, the meter is calibrated and operated similarly to the vacuum-tube voltmeter indicator already described. The circuit conditions make it necessary to have a greater degree of damping in the meter, hence it does not follow variations in program level with the rapidity of the first type.

234. Use of Volume Indicators.—The lack of standardization of volume-indication speed is very confusing in some instances, especially in chain broadcasting. Perhaps its results can best be brought out by means of an illustration.

Suppose that there are two studios A and B. The first is equipped with a slow meter, M_1 , the second with a meter M_2 , which responds quickly. We will say that the receiving point (such as the transmitter) is checking the level from both studios by means of a slow meter, M_3 .

Both studio meters have been adjusted so that they read 0 db for a steady note such as is obtained from an audio-frequency oscillator. A speaker is using studio A. and the gain is adjusted until the peaks caused by his voice register 0 db. As he emphasizes occasional words, his peaks are high compared to his average level.

Suppose now this same speaker is transferred to studio B and the same adjustment made. The checking point will indicate a lower level; how much lower depends upon the amplitude of the speaker's highest peaks and the length of time they last. As the peaks are now practically instantaneous, the studio meter M_2 , having less inertia, gives a truer indication of the height of the transient voltage than does the checking meter M_3 , the input to which must be sustained for an appreciable time before the needle rises to a maximum. In order to get the same response as before on the checking meter M_3 , it may be necessary to raise the gain until the live meter M_2 reads +6 db on peaks. The program will then have the same level as in the first place.

Suppose that the next program is a band in studio B. The peak values of such combinations of instruments as a band are usually sustained over appreciably long intervals of time. The level is set as before on meter M_2 , so that the program peaks at +6 db. As the peak sound level is long enough to entirely overcome the inertia of both meters, the effect of the difference in inertia between them is nullified. The checking meter M_3 will then indicate a program level of +6 db. If the position of the meters were reversed in the first case, the conditions would also be reversed. It can be seen that, if the two meters M_1 and M_2 have not exactly the same degree of damping, they will not read alike unless the input is a steady tone. For this reason the type of indicator and its degree of damping should be standardized for a studio system or chain network.

As the volume indicator is used to take frequency-response runs on equipment, and in order properly to indicate program level, its response must be independent of the frequencies of the tones involved. Both types are calibrated to read the voltage across a 500-ohm resistive load, and the calibration will not hold with frequency change unless provision is made to insure such a constant value of resistance across the input. The input impedance of the device itself is kept high (5 to 20,000 ohms), so that it takes an inappreciable amount of energy from the circuit.

If a volume indicator is placed directly across a line there may be a discrepancy of many decibels between the readings given for the same actual power level, as the frequency in the line varies from high to low. This is because the impedance of the line varies with the frequency, owing to its distributed capacity. The meter should always be isolated from the line by a suitable pad. When used in this manner the resistance across it is kept practically constant at all frequencies, and the meter

calibration remains accurate. A typical set-up is to have the output transformer work into a pad which feeds the line. The meter is then placed across either termination of the pad. If it is placed across the pad input, a better isolation will be obtained; but the reading will be higher than the actual line level by the value of the pad. The results obtained by placing it at the line end of the pad are usually sufficiently accurate for the purpose. When used to make frequency runs, the meter should always be isolated from the equipment under test by a pad of the proper value.

235. Equalizers and Equalizing. *a. Requirements.*—All untreated long lines carrying programs will deliver at their remote end a reproduction lacking in high frequencies. The longer the line, the greater its

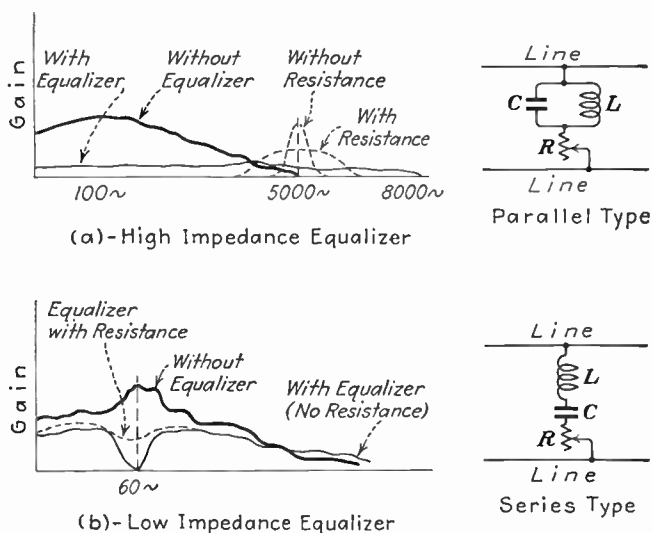


FIG. 243.—Parallel- and series-resonance equalizers.

distributed capacity; and it is this capacity between the conductors which tends to act as a short circuit to the higher program frequencies. Because the reactance of the line decreases as the frequency increases, the higher the note to be transmitted, the greater the attenuation introduced by the line.

To correct this defect, adjustable filters called *equalizers* are inserted in the line. These are set to attenuate (filter) the low notes in exactly the same proportion as the line loses the highs. In this way the loss occasioned at every frequency is equal, and the program reaching the far end of the line is an exact, though attenuated, reproduction of that impressed.

Common forms of equalizers and their characteristic curves are shown in Fig. 243. The coil L possesses a low ohmic resistance (about

3 ohms) and is resonant with the condenser at 5,000 cycles. Some equalizers designed for very long lines, meant for speech only, resonate at 3,000 cycles. In series with the resonant circuit is a decade resistance box R . The whole unit is placed across the line as shown.

The parallel-resonant circuit, (a), offers an extremely high impedance to frequencies around 5,000 cycles, and very little energy at the higher frequencies is lost through this circuit. As the frequency drops below the resonant value, the impedance of the tuned circuit becomes low and allows the lower frequencies to be shunted through it and around the line termination to an extent governed by the setting of the decade box R . By adjustment of this resistance the equalizer can be made to control within its limits the frequency response of any line. If the line loss is greater than about 12 db at 1,000 cycles, additional equalization will probably be necessary. Equalizers may then be used in parallel or special types having a greater low-frequency loss may be employed. It must be remembered that the greater the equalization, the greater the overall attenuation.

b. Operation.—When a parallel-resonant circuit is connected across a line as in Fig. 243(a), it represents a high-impedance load across it. Consequently a maximum voltage will be developed across it at the particular frequency to which it is resonant. Let us assume for example that an audio-frequency line with considerable length attenuates most of the higher range of frequencies above 5,000 cycles owing to the distributed capacity in the line. If the equalizer is adjusted to say 5,000 cycles, the feeble energy developed across it at the 5,000-cycle frequency will be raised by the e.m.f. developed across the high impedance. Thus, the frequency characteristic will be raised at the higher frequencies and decreased at the lower frequencies. Hence, the overall frequency will be flattened out to effect a straighter characteristic over a wider band. This condition is illustrated in the curves in Fig. 243(a). It will be seen from the dotted curves that the series resistance serves to flatten out the characteristic at the resonant frequency.

The series-resonant equalizer of Fig. 243(b) will serve to function in a way just opposite to that of the parallel circuit since, when it is adjusted to resonance, it functions as a low impedance or attenuating device. That is to say, when the reactances X_L and X_C are equal, the effective reactance will be zero, since the reactances cancel each other and consequently no e.m.f. will be developed across it. Hence only the series resistance will be effective. From this it may be seen that if it is desired to attenuate a certain frequency a series-resonant equalizer must be used.

The curves in Fig. 243(b) illustrate the effect of attenuation upon a certain frequency by adjusting L and C to equal reactance values. The amount of attenuation effected by L and C is governed by the resistance R , the amount of attenuation varying with the value of R in the circuit to the right.

c. Use.—An equalizer may be used at any point in the line, but it is best practice to insert it at the receiving end. It is at this point that the quality of reception will be judged, and emergency conditions which may require readjustment of the equalization must in most cases be met. There are other important considerations. With the filter (equalizer) at the terminating end, any low-frequency noises picked up by the line will be attenuated by the equalizer. The longer the line, the more noise there is apt to be picked up; but, as the equalizer resistance will be correspondingly low, the

greater will be its effectiveness in decreasing the noise level. If the low frequencies are attenuated at the sending end, the line noises will arrive at the receiving end proportionately greater in respect to the program than they would be with the filter at the terminating end.

If a filter circuit used with zero or low resistance is located at the point of origin, it offers a very low impedance at the lower frequencies which impedance is directly across the output equipment. Additional precautions are then necessary to isolate sufficiently the volume indicator and amplifier output to insure that the frequency response of both units remains unaffected.

236. Frequency Runs.—In order to equalize a line, an audio-frequency current of constant value, but varying in frequency from that of the lowest tone to that of the highest to be transmitted, is impressed upon the circuit. The filter is then adjusted until the volume indicator at the remote end gives the same reading, within the prescribed tolerances, at each frequency.

The source of steady tone is an audio-frequency oscillator. In order to get accurate results it is important that the harmonic content of the impressed frequency be negligible in value. The oscillator output is amplified and impressed on the line through an isolating pad. It is very important that the impedance across the input of the volume indicator be the same regardless of frequency. For this reason the instrument should be bridged across the amplifier side of the pad. Runs made without such precautions as to the volume-indicator termination may be a number of decibels off at either the low or high end, depending upon the termination.

At the receiving end, the line output, after being equalized, is passed into a suitable line-terminating coil, the output of which is passed through a pad. The power level at the pad output may be measured directly on a sensitive thermocouple meter calibrated directly in decibels. More often it is subject to further amplification by the equipment to be used in conjunction with that line, or failing this, by an amplifier having known characteristics, the output of which is fed to a properly terminated volume indicator.

Preliminary adjustments of the equalizer resistance at three or four of the most important frequencies, perhaps at 100, 1,000, 3,000, and 5,000 cycles, are made to determine the approximate value of the resistance required. Finer adjustments are then made until the response of the line is sufficiently flat over the required frequency range.

Until recently lines for remote pick-ups have not been equalized for frequencies over 5,000 cycles. There are now available, and in use at the better stations, filters which can flatten moderately long lines up to 8,000 cycles. Of course, if necessary, lines can be equalized over a range of frequencies wide enough to transmit a broader frequency spectrum. The lines between studio and transmitter are usually made good

from 30 up to 8,000 or 10,000 cycles, and in some instances up to 15,000 cycles.

When broadcasting from remote points there are usually at least two separate telephone lines installed. One is a *radio line* to carry the program, and the other is a *talking line* used for communication between the studio and the remote point. To equalize such lines, a regular remote-control amplifier is sent to the remote terminating point. Then the output of an audio oscillator is impressed upon the talking line at the studio end. These oscillations are amplified by the remote amplifier and impressed upon the radio line at the remote end. Then, by adjusting the gain of the remote amplifier, a given level is secured. At the studio end, the radio line is then equalized, readings being taken on the equipment which will be normally associated with it. After this process is completed, the lines are reversed and the talking line is equalized in the same fashion. Thus, should a breakdown occur on the radio line, the talking line may be substituted without the quality of the reproduction suffering.

If the lines are long, it is necessary to insert *booster amplifiers* at intervals. This keeps the program level well above the line noise at all points. For high-quality transmission such boosters should be inserted every 10 to 20 or more miles, depending a great deal upon the natural noise level in the cable and the amount of equalization loss entailed. The higher the frequencies equalized for, the greater the loss in program level and the closer the amplifiers. Such units are supplied by the telephone company as part of its service.

It should be mentioned that radio and talking lines do not go through a switch-board operator but are permanently connected between the studio and the remote pick-up point. The radio lines should be tested daily with an oscillator to determine their frequency and gain characteristics. Occasionally the intermediate boosters introduce distortion due to defective tubes. This must be reported to the telephone company.

237. Remote-control Equipment.—This apparatus is similar to studio equipment except that it is constructed in such a way as to achieve utmost portability. It consists of an amplifier, mixer, and volume indicator and associate essentials. Low-drain tubes are usually used with a dry-battery power supply, although airplane-type light storage batteries are sometimes used for filament supply.

Because of their high output and portability, carbon microphones are generally used, though for special occasions or otherwise difficult pick-ups, some other type of microphone may be employed. As the amplifier is designed primarily for use with the carbon microphones, an extra preamplifier is usually necessary with other types. The exact design of such equipment is highly individual and varies as the requirements of the stations employing it.

Smaller stations, in the interests of economy, do not always send an operator to each remote broadcast. If a certain dance band is to be picked up at regular intervals, an amplifier and microphone are left set up at that point. In order to go on the air the announcer turns on the filament switch and, after making tests with the studio to assure all concerned that the system is functioning, stands by for his cue to proceed with the broadcast. This cue is usually delivered over the talking line,

either verbally or by means of signaling with the calling bell. With such a system no volume indicator is required, nor is a mixer usually used.

The equipment must be capable of withstanding great changes in program level without distortion. The output tube should have a maximum undistorted output of at least +14 db, and the gain control should precede the first tube in the amplifier. The output of the amplifier is adjusted so that the program level on the line is not normally higher than necessary to cover line noise. When intelligently used and given adequate maintenance, such set-ups are entirely satisfactory in their results.

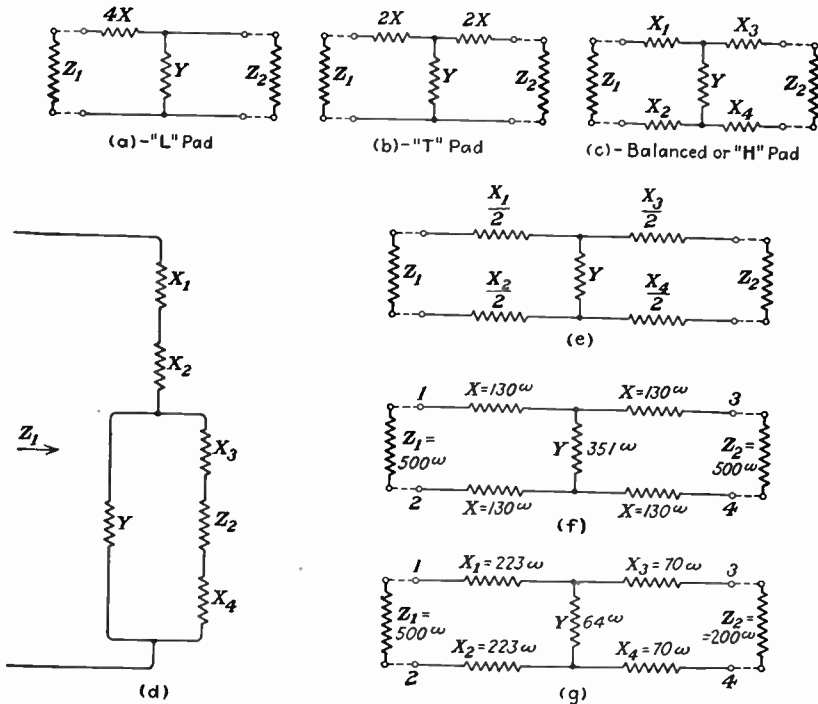


FIG. 244.—Various types of attenuation networks.

238. Attenuation Networks.—These networks are used when it is desired to attenuate a transmission, isolate equipment, change transmission impedance, or balance a line.

There are various forms of attenuation networks, the most common being the *L*, *T* and *H* pads. The names are derived from the arrangement of the resistors comprising them as illustrated in Fig. 244 (a), (b), and (c), respectively.

The *L* or *T* pad is used when line balance is not a factor. The *H* type, which is the one most commonly used in broadcasting, not only does not disturb the balance of the conductor impedances, but actually

serves to equalize them by decreasing the ratio of unbalance if it is present. The greater the value of the series elements, the more pronounced this effect.

In order to prevent reflections from taking place, the input and output impedances of the pad must match that of their terminations. A simple analogy of electrical reflection is presented by light which when striking a surface is partially reflected and partly absorbed. A mirror absorbs very little light and reflects a good deal, a truly dark object absorbs all the light and reflects none. In an electric circuit, that portion of the energy not absorbed by the termination impedance will be reflected back to the source, just as is the light striking a partially reflecting surface. At the source, reflection will again take place, and the energy will travel back and forth between the source and the termination until it gradually expends itself. If the impedance of the termination matches that of the source, all the energy will be absorbed by it and none reflected. This is one of the reasons why impedance matching is so important, particularly on long lines or involved systems.

By means of the following equations the proper values of the various resistance units of a pad may be calculated.

FORMULAS APPLYING TO PADS WORKING BETWEEN EQUAL IMPEDANCES

Note: The letters X , Y , and Z as used in the following equations refer to the resistance branches so lettered in the pad circuits in Fig. 244.

For the H Pad [see diagram (c)]:

$$X = \frac{Z(K-1)}{2(K+1)}$$

but if

$$\frac{K-1}{K+1} = X^K$$

then

$$X = \frac{Z}{2} X^K \quad (167)$$

$$Y = \frac{2ZK}{(K+1)(K-1)} = 2Z \frac{K}{K^2-1}$$

but if

$$\frac{K}{K^2-1} = Y^K$$

then

$$Y = 2ZY^K \quad (167;)$$

Proof:

$$Z = 2\sqrt{X(X+Y)}$$

The constant K is fixed by the attenuation desired and is found in this type of problem by working out the expression, $\text{db} = 20 \log K$.

Table VII which follows gives the values of X^K and Y^K for various values of decibels (*db*).

TABLE VII.—THE CONSTANTS* X^K AND Y^K FOR *H* AND *T* PAD COMPUTATIONS
For pads working between equal impedances only

<i>db</i>	<i>K</i>	X^K	Y^K
1	1.122	0.057	4.48
2	1.259	0.115	2.135
3	1.413	0.171	1.438
4	1.585	0.226	1.036
5	1.778	0.280	0.823
6	1.995	0.333	0.669
7	2.24	0.382	0.560
8	2.51	0.430	0.473
9	2.82	0.481	0.405
10	3.16	0.519	0.351
11	3.55	0.560	0.305
12	3.98	0.597	0.268
13	4.47	0.636	0.235
14	5.01	0.667	0.207
15	5.62	0.697	0.184
16	6.31	0.723	0.162
17	7.08	0.752	0.144
18	7.94	0.776	0.127
19	8.91	0.798	0.116
20	10.0	0.818	0.101
21	11.2	0.836	0.090
22	12.6	0.853	0.079
23	14.1	0.867	0.071
24	15.8	0.880	0.063
25	17.8	0.893	0.056
26	20.0	0.904	0.050
27	22.4	0.914	0.044
28	25.1	0.923	0.039
29	28.2	0.931	0.035
30	31.6	0.938	0.031
31	35.5	0.945	0.028
32	39.8	0.950	0.025
33	44.7	0.956	0.022
34	50.1	0.960	0.019
35	56.2	0.965	0.017
36	63.1	0.967	0.015
37	70.8	0.972	0.014
38	79.4	0.975	0.012
39	89.1	0.977	0.011
40	100	0.980	0.010
41	112	0.982	0.0089
42	126	0.984	0.0079
43	141	0.985	0.0070
44	158	0.987	0.0063
45	178	0.988	0.0056
46	200	0.990	0.0050
47	224	0.991	0.0044
48	251	0.992	0.0039
49	282	0.993	0.0035
50	316	0.993	0.0031

* $X^K = \frac{K-1}{K+1}$; $Y^K = \frac{K}{K^2-1}$; $K = \text{antilog } \frac{db}{20}$; $K = \frac{\text{input } E \text{ or } I}{\text{output } E \text{ or } I}$.

$$K = \text{antilog} \frac{\text{db}}{20} \quad (168)$$

As an example a pad having an impedance of 500 to 500 ohms and a loss of 10 db will be solved.

From the above expression

$$K = \text{antilog} \frac{10}{20} = K = \text{antilog} 0.5$$

but

$$\text{antilog} 0.5 = 3.16$$

therefore,

$$K = 3.16$$

Then

$$X = \frac{Z(K-1)}{2(K+1)} = \frac{500 \times 2.16}{2 \times 4.16} = \frac{1,080}{8.32} = 130^\omega$$

or, taking the value of X^K from the table, which is approximately 0.52, the expression becomes

$$X = \frac{Z}{2}(X^K) = 250 \times 0.52 = 130^\omega$$

It is, therefore, quicker to take the value from the table than to work out the fractional expression involving K . Proceeding, we find

$$Y = \frac{2ZK}{(K+1)(K-1)} = 2Z \frac{K}{K^2-1} = 2Z \frac{3.16}{8.98} = 2Z 0.351$$

Therefore,

$$Y = 2Z \times 0.351 = 351^\omega$$

or, taking the value of Y^K from the table, which is 0.351, the expression becomes

$$Y = 2Z \times 0.351 = 351^\omega$$

This pad is shown in diagram (f).

Again it is seen that the table of values reduces the mathematical labor involved. It is seen then that by the aid of the table it is a relatively simple job to compute any value of H or T pad.

It now remains to prove that this result gives the required pad. First, a check of the impedance match will be made. Looking into it from the input, the impedance of the pad should be 500 ohms when terminated by its proper resistance.

By rearranging the circuit diagram a little, the checking process may be facilitated. Thus the H pad, shown in diagram (f), becomes as shown in diagram (d). From the viewpoint of the generator there are two

resistors X_1 and X_2 , having a total value of 260 ohms, in series with a parallel combination consisting of Y which is shunted by $X_3Z_2X_4$, the load resistance being Z_2 . The lettering on drawings (c) and (d) denote equivalent resistors, that is, X_1 in drawing (c) equals X_1 in drawing (d), and so on. The parallel combination has a total resistance of about 240 ohms, as can easily be verified. Thus the total resistance, the combination presents to the source, is 500 ohms. This value is quite accurate enough for the work involved and is probably as accurate as would be obtainable in the actual construction of a pad owing to the tolerances of the commercial resistors used for this purpose.

As the impedances are equal, and the pad symmetrical, there is no necessity to check the output impedance as it must be the same as the input impedance just checked.

An easy way to check the loss is to imagine a convenient value of power as being impressed upon the input terminals and compute from this how much actually reaches the termination.

Thus, in the example just cited, assume that a power of 500 watts is being impressed at the input. This will produce a current of 1 amp., as the pad presents a 500-ohm resistance and $W = I^2R$. One ampere then flows through the parallel combination shown in diagram (d). This current will divide itself between the two paths in proportion to their conductances; that is, inversely proportional to their resistances. Thus, if for example we let $Y = 351$ and $X_3Z_2X_4 = 760$ ohms, then

$$I_{X_3Z_2X_4} : I_Y = 351 : 760 \text{ and } I_{X_3Z_2X_4} = \text{total } I \left[\frac{351}{351 + 760} \right] = 0.310 \text{ amp.}$$

As the value of current in all parts of a series circuit is constant, 0.310 amp. will flow through Z_2 . The power delivered will then be

$$I^2R = 0.1 \times 500 = 50 \text{ watts,}$$

and the loss in decibels may be computed as follows:

$$\text{db} = 10 \log \frac{P_1}{P_2} = 10 \log \frac{500}{50} = 10 \log 10$$

but $\log 10 = 1$, therefore $\text{db} = 10$. Thus the pad has the required termination.

For the T Pad [see diagram (b)]: $X = Z(X^k)$ and $Y = 2Z(Y^k)$ (X^k and Y^k taken from table.)

FORMULAS APPLYING WHEN THE INPUT AND OUTPUT IMPEDANCES ARE NOT EQUAL

Then,

$$\text{db} = 20 \log \frac{I_1 \sqrt{R_1}}{I_2 \sqrt{R_2}} \quad (169)$$

A formula for an H pad to work between uneven impedances follows [the symbols are as illustrated in diagram (e)]:

$$Y = \frac{2K\sqrt{Z_1 \times Z_2}}{K^2 - 1} \quad (170)$$

$$X_{1,2} = Z_1 - (Y - K) \quad (171)$$

$$X_{3,4} = Z_2 - \left[Y - \left(K \times \frac{Z_2}{Z_1} \right) \right] \quad (172)$$

The constant K has a value depending upon the attenuation and is found by working out the expression $\text{db} = 20 \log K$,

$$K = \text{antilog} \frac{\text{db}}{20}$$

Let us calculate an H pad to work between 500 and 200 ohms and having a loss of 20 db. The finished pad is shown in diagram (g). Note that the X values must be divided by 2 as per diagram (e).

Solving:

$$K = \text{antilog} \frac{\text{db}}{20}; \quad K = \text{antilog} \frac{20}{20}; \quad K = \text{antilog} 1; \quad K = 10$$

$$Y = \frac{2 \times 10 \sqrt{500 \times 200}}{K^2 - 1} = \frac{6,320}{99} = 64$$

$$X_{1,2} = 500 - (64 - 10) = 500 - 54 = 446, 446 \div 2 = 223$$

$$X_{3,4} = 200 - [64 - (10 \times 0.4)] = 200 - 60 = 140, 140 \div 2 = 70$$

This may be checked for impedance in the same manner as the previous pad. It should be checked from both terminations, however, as it works between uneven impedances.

The check for proper attenuation is accomplished in the same manner as the previous pad.

It will be noted that it is impossible to construct such a pad to have *any* loss or to work between *any* values of impedances, as to accomplish this negative values of resistance would be necessary.

The isolation effect of a pad upon the equipment it is used with is most important. Let us take for example an amplifier having a 500-ohm output and used between remote points and the studios. The output transformer of the amplifier is designed to work into a load of 500 ohms. If the load is very much decreased, the primary inductance of the transformer will be materially lessened with a resultant loss of the lower frequencies. If the output amplifier transformer is terminated directly by the line it is working into, a different impedance for each frequency will result, owing to the difference in their lengths and distributed capacities. Hence the frequency response of the amplifier will change

with the termination. If, instead, it is terminated in a pad having a sufficiently high loss, this will isolate it from the line enough to guarantee the same characteristics regardless of its location or the pad termination.

Let us find the isolation afforded by a 500-to-500 ohm H pad of 10-db loss and compare it to that offered by a similar network having a 40-db loss. The values for a 10-db pad are $X = 130$ and $Y = 351$. Referring to diagram (c) and substituting these values we find that, under the extreme circumstances of zero termination of the pad output, the resistance across the transformer would be 610 ohms instead of 500, a variation of about 22 per cent. In the other extreme possibility of a short circuit across the pad output, the resistance across the transformer would be 410 ohms instead of 500, a variation of 18 per cent. Thus, using a 10-db pad, the greatest possible variation at the transformer does not much exceed 100 ohms, while if the transformer were used without the pad it would range from zero to an infinite value.

The values for the 40-db pad would be $X = 245$ and $Y = 10$. This pad then represents a resistance of 499.9 ohms when properly terminated. On open circuit at its output this changes to 500 ohms, or a maximum change of about 0.02 per cent. Such a pad, therefore, offers practically perfect isolation against impedance changes affecting the equipment with which it is used.

NOTE: As supplementary reading to this chapter the student is referred to the following reference: BALLANTINE, STUART; High Quality Broadcast Transmission and Reception, *I. R. E. Proc.* May, 1934.

CHAPTER IX

BROADCAST TRANSMITTERS

The term *broadcast station*, as defined by the Federal licensing authority, is a station used for the dissemination of radio-telephone emissions intended to be received by the public. Such a station operates on an assigned carrier frequency within the broadcast band, namely, that band of frequencies extending from 500 to 1,600 kilocycles.

239. The broadcast station consists essentially of a studio, a radio transmitter, and a radiating system. The layout for a complete station is given in the block diagram of Fig. 245. Here it is seen that the studio is located at a distance from the radio transmitter. This is now the general practice and is found to have many advantages. As the studio

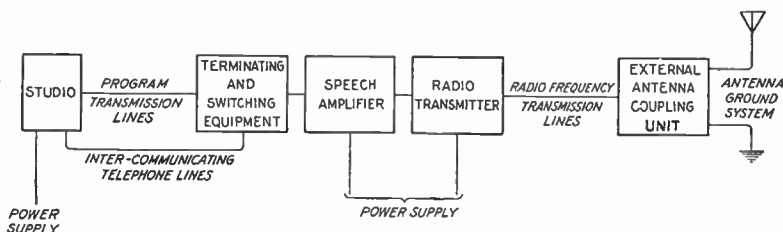


FIG. 245.—Block diagram showing arrangement of broadcast-studio and transmitter equipment.

is the point at which the programs originate and is the central point of control and distribution of lines connecting with various sources of program throughout a city, it is desirable to have the studio located conveniently in the city.

The radio transmitter, however, is best located outside of the built-up areas, where the radiating system is not affected by large buildings and other steel structures. A better field-intensity pattern and overall coverage are then obtained.

The studio and the radio transmitter are interconnected by means of program-transmission lines, which are special telephone lines designed to have a uniform-frequency transmission characteristic from approximately 60 to 6,000 cycles. Two or more independent lines are usually provided to insure uninterrupted service. There are also private intercommunicating telephone lines to permit the operating personnel of the studio and radio transmitter to be in immediate contact whenever necessary. The program lines are terminated in suitable switching and attenuating apparatus which permits the selection of a program line and also controls

the input level to the speech amplifier. This amplifier brings the low program level coming in on the lines up to the value necessary to feed into the radio transmitter where the carrier is modulated. The output of the radio transmitter is then fed into a radio-frequency transmission line which transmits the modulated carrier power to the antenna-coupling unit. This unit is essentially a transformer coupling the antenna-ground system to the transmission line with a proper matching of impedances and at the same time maintaining a proper balance of the transmission line. The antenna and ground constitute the radiating system of the broadcast station and may be in any one of the several conventional forms now in use.

The box layout shown in Fig. 245 may, of course, be varied to meet different conditions. For example, the studio and radio transmitter may be located together, thereby eliminating the expense of high-quality program-transmission lines. On the other hand, the radio transmitter may also be operated directly into the antenna-ground system, thus eliminating the radio-frequency transmission line and coupling unit with their attendant losses. In general, broadcast stations of less than 5-kw. output do not employ a transmission line or external coupling unit.

The studio with all its associated equipment was discussed in the preceding chapter. The switching apparatus and speech amplifier at the radio transmitter are similar to studio equipment, the only difference being in their application. A detailed discussion of the operation of the radio transmitter and all its associated apparatus follows.

240. Definition and Rating of the Transmitter.—By *transmitter* is meant that apparatus wherein a fundamental radio-frequency wave is generated, amplified, modulated (further amplified in some cases), and finally delivered into a radiating system. By *radio-transmitting equipment* is meant the radio transmitter and all apparatus directly associated with it, including the speech-input amplifier, monitors, radio-frequency transmission line, antenna-coupling unit, and all power-supply and control circuits.

Radio-transmitting equipments may be classified in different ways such as low- or high-power; motor-generator-operated; direct-current-rectifier- and motor-generator-operated; or all-alternating-current-operated; low- or high-level modulation; direct- or transmission-line-fed antenna-ground system; etc.

The power rating of a radio transmitter is purely a relative figure and the dividing line may be chosen arbitrarily. At present carrier powers of less than 5 kw. may be called *low* power, and 5 kw. or more, *high* power. The maximum rated carrier power of a broadcast transmitter is determined by the number and types of tubes in the last radio stage and by the type of modulation employed.

For high-level modulation the maximum rated carrier power of a transmitter is the same as the total installed tube-power capacity of the last radio stage. For example, if two UV-861 tubes are employed, this power is 1,000 watts, because these tubes are given a fixed power rating of 500 watts each by the manufacturer when operated under the above conditions with proper plate, filament, and bias voltages and output impedance.

When low-level modulation is used, the maximum rated carrier power is one-fourth the total installed tube-power capacity of the last radio stage. In this case, if two UV-861 tubes are used in the last radio stage, the rated power would be 250 watts.

If grid-bias modulation is employed in the last radio stage, the maximum rated carrier power is the same as the total installed tube-power capacity of that stage, as determined by the tube-power rating for this type of modulation.

It must be remembered that the three above conditions deal with maximum rated carrier power of the transmitter, not with the actual authorized operating power of the broadcast station. In general, the two powers are the same, but sometimes the latter is less than the former, that is to say, a station may be authorized to operate on 750 or 500 watts while it is really capable of 1,000 watts output. However, a broadcast station with an authorized carrier-wave power over 5,000 watts is not permitted to install a transmitter with maximum rated power exceeding twice the operating power.

241. The operating power of a broadcast station (which is the same as the licensed or authorized power) may be determined by one of two methods, direct or indirect measurement. The direct method requires the resistance value of the antenna at the operating frequency as measured with carefully calibrated instruments in approved circuit arrangements. The power is then this value of resistance multiplied by the square of the antenna current in amperes, as measured by an accurate radio-frequency ammeter inserted in the antenna-ground system at the point where the resistance was measured. Another method involves field-intensity measurements for computing the radiated power, but this system is not in general use for broadcast-station installations.

The indirect-measurement method of determining operating power is the accepted standard and is always resorted to when the antenna resistance is unknown or has been changed owing to modification of the antenna-ground system, or when any other factor arises which makes the direct-measurement method of doubtful accuracy. Actually the operating power whether determined by the direct or the indirect method will be the same, providing all measurements and calculations are reasonably accurate and the transmitter is operated normally throughout at its rated power output.

The operating power is determined by the indirect method from the plate-input power of the last radio stage and is equal to the product of the plate voltage, the total plate current in this stage, and a factor F ,¹ which is determined by the type of modulation used and the maximum rated carrier power of the transmitter.

For high-level modulation the factor F is 0.5 for 100 watts output; 0.6 for 250 to 1,000 watts output; and 0.65 for 2,500 to 50,000 watts output. For example, a transmitter has a rated carrier power of 1,000 watts, then the factor is 0.6; if the plate voltage is 3,000 volts and the total plate current is 0.555 amp., then the operating power is $3,000 \times 0.555 \times 0.6 = 1,000$ watts. This is merely another way of stating that the plate-power efficiency is 60 per cent, but it is a reasonably accurate check on the operating power of the transmitter when the direct-measurement method cannot be used, providing the operating constants, such as plate voltage, plate current, radio-frequency input voltage, and output impedance, are all properly adjusted in the last radio stage.

With low-level modulation the factor F is 0.33 for transmitters capable of full 100 per cent modulation. For transmitters which are only capable of 75 to 85 per cent modulation the factor is 0.40, but as all new and modern equipment is designed for full 100 per cent modulation the lower values will be neglected. For example, if an operating power of 1,000 watts and a plate voltage of 3,000 volts are used, then the plate-input power is $1,000/0.333 = 3,000$ watts. Therefore, the total plate current must be $W/E = 3,000/3,000 = 1.0$ amp.

Where grid-bias modulation is employed in the last radio stage, the factor F is 0.22 for transmitters capable of 100 per cent modulation. Assume once again a 3,000-volt plate potential and a plate current of 0.151 amp. This gives an operating power of $3,000 \times 0.151 \times 0.22 = 100$ watts.

These various figures of the percentage efficiency of the plate-input powers are an indication of the optimum operating conditions for the particular type of tube and method of modulation employed in the radio transmitter. That is, for a certain type of modulation it has been found that the best operating conditions will result in obtaining a certain plate-power efficiency and that any great deviation from that efficiency is caused by incorrect operating voltages or impedances with the result that the modulated carrier is distorted to some greater or smaller degree.

Although modulation has been referred to several times, and the factor F is largely determined by the type of modulation used, it must be remembered that all measurements and calculations of operating power, by either the direct or indirect method, must be made only with the carrier unmodulated.

¹ The values for F are given in Sec. 245 (17).

If the operating power of the broadcast station is less than the maximum rated carrier power and the indirect-measurement method is used for determining the operating power, the above stated values of the factor F still must be adhered to, even though the output impedance and input radio-frequency voltage have to be readjusted in order to maintain the plate current at such a value that the

$$\text{Output power} = \text{plate volts} \times \text{plate amperes} \times F$$

Thus,

$$P_o = E_p \times I_p \times F \quad (173)$$

A broadcast station may not at any time have an operating power greater than the maximum rated carrier power. A station may not be operated with a greater percentage of modulation than that for which the transmitter was designed. A station is not permitted to operate at any other than the assigned power output without specific authorization from the Federal licensing authority.

The radio transmitter is really the heart of the broadcast station, for it is here that the medium for conveying a program, speech, music, or other intelligence from the sender to the receiver, that is, the fundamental radio-frequency wave, is generated, amplified, and modulated with a certain program and delivered to the radiating system. Although a transmitter may be designed electrically and mechanically in hundreds of different styles and layouts, still the basic principles of generation, amplification, and modulation must be strictly adhered to (see Chap. IV). Figure 104(c) is a block diagram giving a possible arrangement of a transmitter employing low-level modulation. We shall begin by studying the oscillator, its requirements, and operation.

242. Frequency Control.—Due to the congestion of assigned frequencies in the broadcast band it has been found necessary to demand the maintenance of each station on its particular frequency in order to reduce interference among stations in radio receivers. Early transmitters employed a simple vacuum-tube oscillator [see Fig. 104(a)], consisting of a tuned output circuit with sufficient feed back of energy from the output to the input circuits to cause sustained or continuous oscillations in the former. But it was found that many factors affected the stability of the oscillator, such as plate and filament voltages, temperatures, circuit constants which varied with aging, tube constants, and so on.

The advent of the crystal controlled oscillator resulted in the solution of the problem of stabilization to a certain degree. The frequency is still affected by the above mentioned factors but only very slightly compared to the stabilizing effect of the quartz crystal. However, the resonant frequency of the crystal may be changed by the temperature and by the value of shunting capacity across it. Once the circuit is

assembled, the capacity may be considered constant. The temperature range must be kept within very narrow limits by special housing and by automatic-control circuits.

In practice the oscillator must be designed so that the operating frequency of the broadcast station will be maintained within 50 cycles of the assigned frequency. Furthermore a means must be provided for checking this frequency. This is usually accomplished by having another crystal-controlled oscillator capable of at least the same precision as the oscillator in the transmitter but operating entirely independently of the transmitter. The frequency of the transmitter is then compared with that of a separate oscillator or *frequency monitor* as it is called. True, this is merely a comparison of two similar pieces of equipment, one with another, but in general both oscillators will not fail at the same time, and, since the frequency monitor is operating under nearly constant conditions with no load variations, it will therefore serve as a fairly reliable checking instrument (see Sec. 247).

243. Western Electric Broadcast Transmitter.¹—A schematic circuit arrangement of a modern high-quality broadcast transmitter together with its complete power supply and relay system is shown in Fig. 246. This transmitter incorporates the most modern advances in the broadcast-engineering field. The entire transmitter is alternating-current operated and does not use any motor generators for plate, bias, or filament power. All power circuits are automatically operated by relays and are suitably protected by overload circuit breakers and fuses. Door switches are provided for opening the high-voltage plate-supply circuit to protect against possible injury to the operator when adjustments or replacements are made necessary.

The complete transmitter consists of two units, namely, the 100-watt section and the power amplifier, called the No. 12-A and No. 71-A units, respectively. Both of these units coupled together form a 1-kw. transmitter, or the 100-watt section may be operated individually as a complete 100-watt transmitter.

The theoretical operation of the transmitter follows the principles laid down in Chap. IV, Transmitting Circuit Principles. In this chapter the practical operation of this broadcast transmitter will be treated in detail. The six-tube full-wave three-phase rectifier which supplies power for this transmitter is treated from a theoretical standpoint in Chap. XIII, Rectifier Units, later in the book, and should be studied carefully.

¹ The descriptive material and diagrams in this and the following sections having to do with Western Electric apparatus have been adapted to the purposes of this text partly through actual practical experience with the equipment described and partly from copyrighted instruction books supplied to the authors for this purpose through the courtesy of the Western Electric Company.

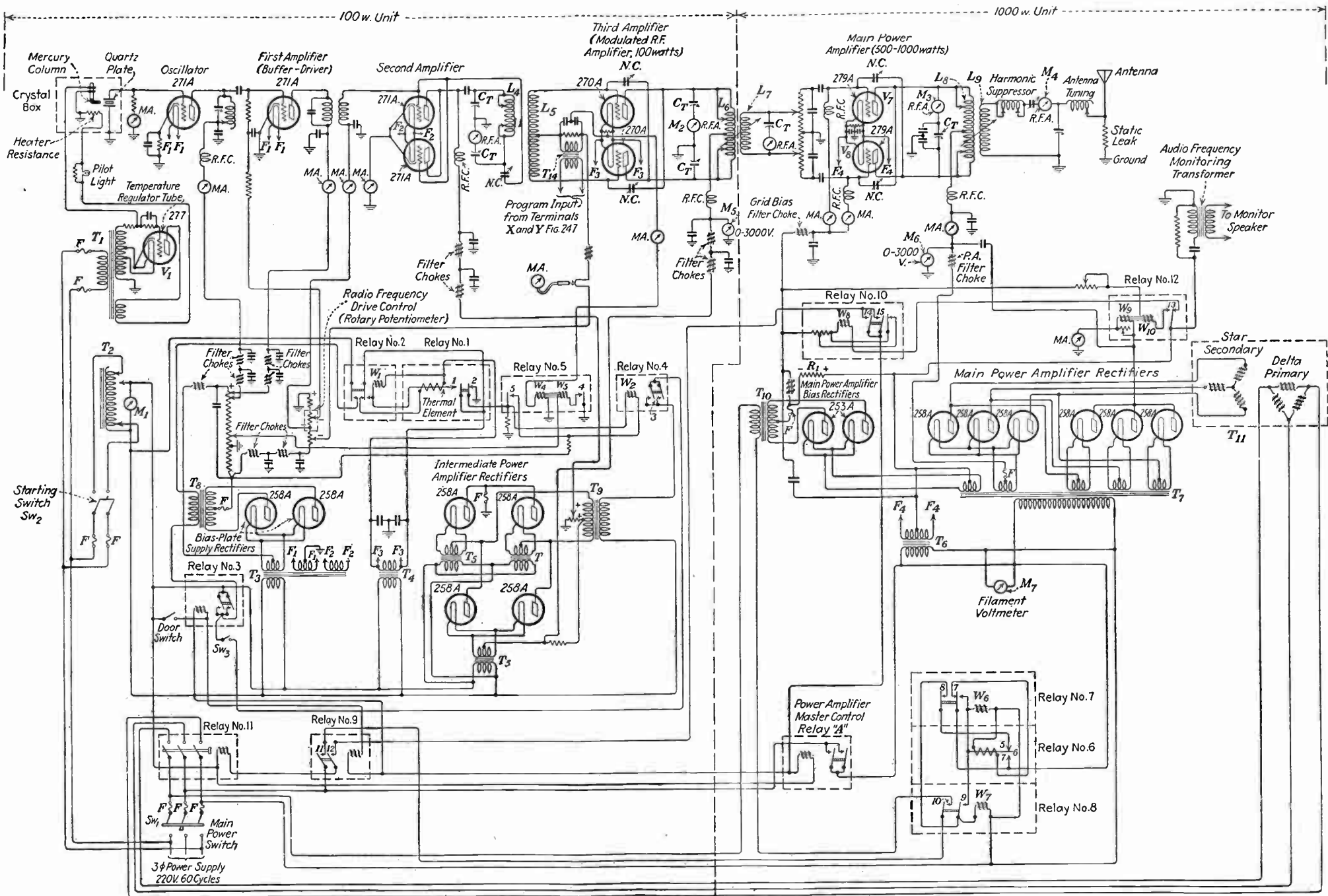


FIG. 246.—Western Electric 1 Kw. Broadcast Transmitter.

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244. Relay Operation.—The practical operation of the power-control and protection circuits of the entire transmitter when the main power switch Sw_1 is closed, is described as follows.

1. The switch Sw_1 connects a three-phase 220-volt 60-cycle source to the various circuits of the 12-A and 71-A units and is closed when the transmitter is in operation. These circuits, with the exception of the transformer T_1 , will, however, not be energized by the closing of this switch. The transformer T_1 must be constantly energized in order to operate the temperature-control chamber of the crystal box to maintain a constant-operating frequency. The primary winding is, therefore, permanently connected across the alternating-current line. If this switch is left open for a considerable time the temperature in the crystal chamber will be altered, and a frequency deviation of several hundred cycles may result. This transformer supplies the filament, plate, and bias voltages for the argon regulator tube V_1 . The operation of this circuit will be explained later.

2. The next switch to be closed is Sw_2 . This supplies an alternating-current voltage across an autotransformer T_2 . A tapped portion of this transformer connects to a rotary switch for the purpose of selecting the proper alternating-current voltage to be delivered to the operating circuits. This provides the proper flexibility necessary to maintain the desired voltage to the load circuits to compensate for line-voltage changes. This voltage must be maintained at 220 volts as indicated by the alternating-current line voltmeter M_1 . Any variation in the supply-line voltage may then be regulated by the rotary switch to the required value.

The autotransformer voltage excites the primary windings of all the filament-lighting transformers in the 12-A unit, T_3 , T_4 , and T_5 and the master-control relay A in the 71-A amplifier unit, which closes and excites the primary windings of the filament transformers T_6 and T_7 for the power amplifier and rectifier tubes, respectively.

3. Connected across the secondary winding of the filament transformer T_4 is a resistance unit which becomes heated when the winding is energized. This unit (and its associated contacts) is known as a *thermal or heater-element time-delay relay*, relay 1. This resistance is located very close to an alloy strip (thermal element) which, when heated, begins to expand owing to the effect of heat on the metal. As this metal strip expands sufficiently it finally makes contact with the terminal 1 on the unit and energizes another relay coil winding W_1 of relay 2. The magnetized winding of this coil attracts two metal contact strips which, when closed, open up the connection to the heater unit of relay 1 and close the circuit leading to the primary winding of the transformer T_8 . This transformer supplies the plate voltage to the oscillator and buffer-amplifier stages only and also supplies the negative bias to the grids of the first amplifier and modulating radio-frequency amplifier stages. It is important to note, however, that relay 2 cannot be actuated unless relay 3 is also closed. The latter can only be operated when all door switches are closed by tightly locking all compartment doors.

Now, since the heating resistance unit of relay 1 is open-circuited, the heating effect upon the metal strip will be decreased and consequently will cause the strip to return back to its normal position. This closes the back contacts 2 and completes the circuit through the relay 4, winding W_2 . This closes the contact 3 on this relay and excites the primary winding of the plate-voltage transformer T_9 . It is assumed, of course, that the plate-supply switch Sw_3 and the door-switch relay 3 have been previously closed.

4. The overload relay 5 in the 12-A unit of the transmitter consists of two coils, namely, the overload-coil winding W_4 and the operating winding W_5 . The overload coil is connected in series with the center-tapped filament return lead to ground which

completes the plate-current circuit for the modulated radio-frequency amplifier tubes. If an abnormal flow of plate current should result in these tubes due to excessive modulation peaks or line-voltage surges, the winding W_4 of relay 5 will be energized sufficiently to draw over the contactors and close the circuits 4 and 5. This action will short-circuit the winding of relay 4 thereby demagnetizing it and releasing the contacts 3 on the relay 4. Thus the primary winding of the plate-supply transformer T_9 will be opened and the high voltage to the modulated amplifier tubes will be disconnected.

5. A small-overload reset push-button (not shown) is provided to open the winding W_5 so that the short circuit will be removed from the winding W_2 on relay 4 and thereby reestablish the plate voltage to the modulated amplifier tubes.

6. The relays in the 71-A unit operate in a similar manner to those in the 12-A unit with the exception that they are designed to carry relatively larger values of currents.

When the master control relay A is closed, the primary winding of the filament transformers T_6 and T_7 are excited as stated before. At the same time, however, another thermal time-delay relay 6 is heated, since it is connected across the primary windings of the two filament transformers T_6 and T_7 . This results in the expansion of the thermal element and the closing of contacts 6 and 7; and this allows current to flow through the winding W_6 which energizes relay 7. When this relay is energized, the contact 7 is opened and the heater winding of relay 6 is broken, thus allowing the thermal element to contract and reestablish connections 5 and 6 on relay 6. At the same time, however, the contact 8 is closed which allows a current to flow through the winding W_7 . This magnetizes relay 8 and closes the contacts 9 and 10. The closing of these contacts allows an alternating current to flow through the primary winding of the transformer T_{10} and establish a voltage between the two plates of the associated mercury-vapor rectifier tubes. This results in a current flow through the resistance R_1 which develops a voltage drop across it. This drop is used to supply the bias of -275 volts to the grids of the two power-amplifier tubes V_7 and V_8 . This is, of course, provided that the relay 9 is also closed by having all door switches locked. Otherwise contact 11 will be open and no current will pass through T_{10} .

7. As soon as the grid-bias voltage is developed across the resistance R_1 , a current will pass through the grid-bias relay 10 because it is connected directly across the resistor through contacts 13. This actuates the relay winding W_8 by magnetizing the iron core and closing the contacts 14 and 15. In series with these contacts is a large solenoid contactor, relay 11, which becomes energized and draws over three contactors to close the three-phase alternating-current supply to the high-voltage transformer T_{11} . This action supplies the alternating-current high voltage to the plates of the main power-amplifier, mercury-vapor rectifier tubes, where it is then rectified to the desired direct-current potential of 3,000 volts for the plate supply.

8. All circuits should now be excited and the proper plate, bias, and filament voltages should be applied. These voltages may now be properly checked by the filament voltmeter M_7 and the direct-current voltmeters M_5 and M_6 . All plate and radio-frequency currents may also then be checked by the various plate milliammeters MA , and by the radio-frequency anammeters RFA , in the respective circuits.

9. In the W. E. installation only one milliammeter is used to obtain the plate- and grid-current readings for several of the low-power stages in the 100-watt unit. This is accomplished by a number of resistance shunts and a rotary-selector switch which transfers the meter into the desired circuit. In Fig. 246 this switch and the associated shunts are omitted, but individual meters are inserted in their proper places to simplify circuit analysis.

10. All plate- and power-supply circuits are suitably protected by fuses and overload circuit breakers as illustrated in the diagram. Two overload relays 5 and

12 are provided in the 100-watt and power-amplifier units to enable the operator quickly to reestablish transmitter operation in the event of a circuit breaker "tripping" owing to excessive modulation peaks, line-voltage surges, or temporary condenser flashovers due to dust accumulation.

11. The overload-coil winding W_9 of relay 12 is connected in series with the negative terminal of the 3,000-volt rectifier and ground. If an excessive value of current flows through this coil the core is magnetized sufficiently to pull the relay arm over and break the contacts 13. This opens the winding W_8 on relay 10 and releases the contact arm of the relay, which breaks contacts 14 and 15 and opens relay 11. This disconnects the three-phase supply to the power transformers and cuts off the high-voltage supply to the power-amplifier tubes. The winding W_{10} on the overload relay 12 serves to hold the arm in this position until it is desired to reestablish the plate voltage.

12. A push-button reset switch (not shown) is connected in series with winding W_{10} to break this circuit when it is desired to release the arm back to its normal position, thereby reestablishing contacts 13, 14, and 15 and again energizing winding W_8 of relay 10.

245. General Operation and Tuning.—To place the 1-kw. transmitter into 500-watt operation at an authorized frequency, it will be necessary to adjust carefully all circuits for the proper load conditions as follows:¹

1. All plate, filament, and bias voltages must be carefully adjusted to their required values in accordance with their class of operation.

The crystal oscillator uses a type 271-A tube and operates with a plate potential of 130 volts at a plate current of 7 ma. and a grid current of 0.3 ma. The first amplifier uses a type 271-A tube and operates with a plate potential of 300 volts and draws a plate current of 6 ma.

The second-amplifier stage consists of two 271-A cathode-heater tubes, in series or parallel. These tubes operate at a plate potential of 350 volts, a plate current of 12 ma., and a grid bias of -75 volts. The third amplifier or modulating-radio-frequency stage uses two 270-A tubes (350 watts each) operating as a class C amplifier at a plate potential of 3,000 volts, a total plate current of 125 ma., and a grid bias of -250 volts. These tubes are adjusted to a 100-watt output.

The final, or power-amplifier, stage uses two type 279-A tubes (rated at 1,200 watts each) operating as a class B amplifier with a plate potential of 3,000 volts, a total plate current of 0.52 amp., and a grid bias of -275 volts.

The filaments of the entire transmitter are alternating-current operated through the medium of single-phase power transformers. The filament-operating voltage of the 271-A tubes is 5 volts; and the 270-A and 279-A tubes are all operated at 10 volts.

The high-voltage rectifier tubes are W.E. type 258-A having a maximum inverse-peak potential of 6,500 volts. The grid-bias rectifier tubes are W.E. type 253-A having a maximum inverse-peak potential of 3,500 volts. All rectifiers are of the mercury-vapor type. Type 253-A tubes are used in this transmitter in the rectifier unit, and type 258-A tubes are used in the plate-supply rectifier units. The filament voltage of both types is 2.5 volts. The peak current of type 253-A is 500 ma.; of type 258-A 1.1 amp.

2. Use the crystal-oscillator box corresponding to the newly assigned frequency. This box is adjusted to within 25 cycles of the assigned frequency by the Western Electric Company. Care must be taken that the mercury column in the box is free to rise. This may be assured by slightly tapping the rear edge of the box on a table.

¹For theory underlying operation and tuning see Secs. 117 and 119.

3. Insert a new argon heater-regulator tube but leave the plate voltage for this unit disconnected until its filament has been heated for several minutes. Then close the plate-voltage supply switch. A pilot light will show that the heater unit in the crystal box is being properly heated. It will take at least 2 hr. before the box temperature is correct and the pilot light goes out.

4. During the 2 hr. interim, calculate the radiation resistance of the antenna by the method explained in Chap. VI on Antennas (Sec. 185).

5. Refer to the tuning-adjustment chart and connect coils, condensers, and links to correspond to the desired frequency.

6. Close all door switches by closing doors, and place the master switch in starting position. Various time-delay relays will close in gradual succession but only after the bias-rectifier tubes are in operation. This prevents any possibility of the plate voltage being applied before the biasing circuit is completed.

7. The first amplifier, or buffer stage, requires no tuning adjustments or neutralization because of its aperiodically tuned plate-circuit transformer. This transformer effectively covers the broadcast frequencies of from 500 to 1,200 kilocycles. A variable resistance controls the bias voltage on the grid of this tube and also is the main radio-frequency drive control for the succeeding amplifier tubes.

8. Prepare the second radio-frequency amplifier for neutralizing by opening the third-amplifier plate-supply switch and closing the second-amplifier plate-supply switch.

Close the high-voltage-supply switch on the 12-A transmitter panel and note the plate-current reading of the milliammeter. Adjust the tuning condensers which are connected across L_4 until the plate milliammeter shows a maximum drop. This is a preliminary tuning adjustment before neutralizing and, should difficulty be encountered in the adjustment, the neutralizing condenser should be shifted slightly. It was assumed that the neutralizing condenser was approximately one-fourth engaged.

The second stage is now ready for neutralizing, but it is advisable to proceed first with the neutralization of the third amplifier to simplify matters.

9. To neutralize the third amplifier open the high-voltage-supply switch and open doors. Set the neutralizing condensers on the third-amplifier stage so that they are a little less than one-half engaged. Insert a thermogalvanometer and thermocouple into the plate mesh circuit C_7L_6 . Close the panel doors and adjust the radio-frequency drive to a minimum position and loosen the magnetic coupling of L_6L_7 . Close the high-voltage-supply switch and begin to vary the third-amplifier tuning condensers across L_6 until the radio-frequency ammeter in this mesh circuit reads maximum. If no reading is obtained, increase the radio-frequency drive. Also vary the mesh tuning condensers. An excessive reading will require an increase in the capacity of the neutralizing condensers until the reading drops to a low value. Continue to increase the radio-frequency drive but keep adjusting the neutralizing condensers so that the mesh current does not exceed $\frac{3}{4}$ amp.

Both the second- and third-amplifier tuning condensers must then be varied until a maximum current is indicated in the radio-frequency mesh-circuit ammeter. Then vary the third-amplifier neutralizing condensers until the radio-frequency ammeter in the mesh circuit reads zero.

10. The second amplifier may now be neutralized by opening the high-voltage-supply switch and reducing the radio-frequency drive to a minimum. Open the plate-supply switch to the second amplifier and close the plate-supply switch to the third amplifier. Close the high-voltage-supply switch and vary the radio-frequency drive together with the neutralizing condenser until the radio-frequency ammeter in the third-amplifier mesh circuit reads about $\frac{3}{4}$ amp. Adjust the second- and third-amplifier tuning condensers until the third-amplifier radio-frequency mesh meter M_2 reads a maximum. Then vary the second-amplifier neutralizing condenser until this

meter reads minimum. The radio-frequency drive should then be increased to a maximum and the second-amplifier neutralizing condenser adjusted until no reading is obtained at the radio-frequency meter in the third-amplifier mesh circuit.

11. The power-amplifier stage may now be neutralized by opening its plate-supply circuit and reducing the magnetic coupling between L_8 and L_9 and increasing the capacity of the power-stage neutralizing condensers to about one-fourth capacity. Apply the plate voltage by closing the high-voltage switch on the 12-A unit. Increase the radio-frequency drive and vary the power-amplifier tuning condenser across L_8 until the radio-frequency ammeter M_3 in the plate mesh $C_T L_8$ reads about $1\frac{1}{2}$ to 2 amp. Increase the capacity of the neutralizing condensers until the reading of the power-amplifier radio-frequency ammeter in the $C_T L_8$ mesh reads at, or very nearly, zero.

12. The entire transmitter may now be tuned for maximum efficiency by adjusting the various amplifier stages to the proper load and resonant conditions. All stages with the exception of the first-amplifier and antenna circuits indicate a resonant condition when a minimum plate current is obtained in the tube whose tuned circuit is being adjusted. Great care should be taken in tuning the second and third amplifier so that they will not be resonant to the second-harmonic frequency of the carrier.

13. No tuning adjustments are necessary in the first-amplifier circuits, since all condensers and coils are of the fixed or untuned variety. The second-amplifier circuit is tuned for a maximum dip in the plate current by varying the condensers C_T across L_4 . The third-amplifier output circuit is then adjusted for the desired input to the power-amplifier tubes.

14. Open the plate-supply voltage of the 71-A unit (power-amplifier stage), and insert the full power-input resistance across the grids of the power-amplifier tubes. Close the plate-supply voltage of the 12-A unit and tune condensers C_T across L_8 until the radio-frequency ammeter in this mesh reads a maximum value. Adjust the radio-frequency drive so that this meter does not exceed 1.5 amp. Vary the coupling of $L_6 L_7$, and adjust the power-amplifier input-control condenser C_T across L_7 until the radio-frequency ammeter in the $C_T L_6$ circuit reads a minimum. Increase the coupling of $L_6 L_7$ and the radio-frequency drive until the radio-frequency ammeter in the $C_T L_7$ circuit reads about 1 amp. and the radio-frequency ammeter in the $C_T L_6$ circuit drops between 0.8 and 1.3 amp. The plate current of the third-amplifier stage at this point should read between 125 and 165 ma. Then adjust C_T across L_6 for a minimum plate-current reading in this stage.

15. Before proceeding with the output tuning of the power-amplifier stage, it will be necessary to adjust the radio-frequency drive so that the grids of the power-amplifier tubes receive the proper excitation. This is accomplished by inserting a radio-frequency ammeter in series with the resistances (2,400 ohms) which are across the power-amplifier grids. Assuming 100-watt excitation, the radio-frequency drive is increased until the radio-frequency ammeter in the resistance circuit reads a little less than 0.22 amp. Thus, $W = I^2 R$ or approximately 100 watts. Accurate power-amplifier grid excitation may thus be obtained for any power up to 1,000 watts in this transmitter.

16. The output circuit of the power amplifier may then be tuned as follows. Adjust the harmonic-suppression coil to the reactance value which will offer a minimum impedance to the desired carrier frequency and a maximum impedance to the second-harmonic frequency. This value is obtained by referring to the calibration chart accompanying the transmitter. Close the power-amplifier plate-supply circuit. The plate current in this circuit should read very nearly 0.3 amp.

Vary the radio-frequency drive until the radio-frequency ammeter in the $C_T L_8$ mesh reads approximately 2 amp. At this point recheck the third-amplifier plate current by varying C_T across L_6 until the plate current in this stage is of minimum value.

Vary C_T across L_s for maximum current in the $C_T L_s$ radio-frequency mesh. Increase the output coupling slightly, and vary the antenna tuning condenser in series with the harmonic-suppression coil until the radio-frequency ammeter in the plate mesh reads a minimum.

Increase the coupling $L_s L_3$, and vary the radio-frequency drive until all meter readings in the power-amplifier output circuit correspond with the required manufacturer's ratings for a given power output.

17. If the antenna resistance was found to be 50 ohms, then the radio-frequency drive should be adjusted to cause the antenna ammeter M_4 to read 3.2 amp., or an input of a little over 500 watts. In accordance with the regulations governing broadcast stations, the operating power of a transmitter may be calculated by the formula $P_o = E_p \times I_p \times F$.

The factor F as determined by the Federal licensing authority is as follows:

A. FACTOR TO BE USED FOR STATIONS USING HIGH-LEVEL MODULATION

Maximum rated carrier power of transmitter, watts	Factor (F) to be used in determining the operating power from the plate-input power
100	0.50
250-1,000	0.60
2,500-50,000	0.65

B. FACTOR TO BE USED FOR STATIONS OF ALL POWERS USING LOW-LEVEL MODULATION

Maximum percentage of modulation	Factor (F) to be used in determining the operating power from the plate-input power
75- 85	0.40
86-100	0.33

C. FACTOR TO BE USED FOR STATIONS OF ALL POWERS USING GRID-BIAS MODULATION IN THE LAST RADIO STAGE

Maximum percentage of modulation	Factor (F) to be used in determining the operating power from the plate-input power
75- 85	0.27
86-100	0.22

18. This transmitter employs the low-level system of modulation and is capable of modulating the carrier frequency between 86 and 100 per cent. The operating power is obtained by the product of $E_p \times I_p \times F$, which for this transmitter is $3,000 \times 0.52 \times 0.33 = 515$ watts (approximately).

The correct plate-current values are obtained by careful adjustments of the plate voltages in the various stages, and to the critical variations of the load constants due to the LC ratios and reflected impedances, as explained in Chap. IV on Transmitting Circuit Principles.

19. When the transmitter has been properly tuned and neutralized as described, the proper monitoring level may be determined for the amount of power to be used.

The complete adjustment of the radio- and audio-frequency circuits are carried on into a dummy-antenna load having a resistance of approximately the same value as that of the antenna. This permits the thorough testing of all the circuits to insure their compliance with the rigid requirements of the licensing authority before connecting the transmitter to the antenna or radiating system, thereby greatly reducing interference during the test-hour periods.

Slight retuning of the output coupling circuits must again be made when the antenna circuit is connected for program radiation.

Although the tank circuits are tuned to resonance when adjusted for a maximum dip in the plate-current reading as indicated by the direct-current milliammeter, the full-power output cannot be secured unless the tuning adjustments are slightly altered. In reality the circuit is not detuned but merely changed from a maximum impedance to a unity power-factor condition. This is particularly important if the circuit is adjusted by the inductive reactance X_L since the resistive component of the inductance L prevents an absolute adjustment to these conditions. If, however, the tank circuit is tuned by a condenser, as in Fig. 246, the capacitive reactance may be more easily adjusted so that the combined reactances are equal, thereby resulting in a unity power-factor condition. This adjustment gives a greater transfer of power because the load-circuit resistance will decrease, thereby reducing the power dissipation, due to the resistance, and giving more useful power output. Modulation and frequency measurements must then be made to insure the proper percentage of modulation and frequency stabilization in accordance with the requirements of the Federal licensing authority.

246. The speech-input equipment associated with this transmitter is the W. E. type 9-A illustrated in Fig. 247. This equipment must be properly adjusted to deliver the desired speech-input level for complete modulation.

If the transmitter is to be operated at the full power rating of 1,000 watts, the speech-input level for 100 per cent modulation must be 10 db. If, on the other hand the transmitter is to be operated at 500 watts, the proper speech-input level must be 7 db. This may be obtained by either decreasing the speech-amplifier output level 3 db, or by operating the amplifier at normal and inserting a loss (pad) of 3 db between the speech amplifier and the transmitter input. If a pad is used to decrease the gain 3 db, the proper input and output impedance match must be obtained to insure maximum efficiency. Since the output of the 9-A speech-amplifying equipment is designed to function into a 500-ohm line, and the line is to terminate into the primary winding of the grid-modulation transformer T_{14} (Fig. 246) which has an input impedance of 1,000 ohms, the attenuating network (pad) must be carefully calculated to conform with these requirements (see Sec. 238 on pad calculations).

The theoretical operation of the speech-amplifier system illustrated in Fig. 247 is primarily the same as that of any high-quality audio-frequency amplifier operating as a class A system. That is to say, all tubes are properly biased to operate at the straight portion of the $I_p - E_c$ characteristic curve so that the output-wave shape in each of the plate circuits is a symmetrical reproduction of the input grid voltages. The operations of all component parts and circuits are identical with those explained in other sections. However, in order that a complete picture of the entire system as one unit may be clearly understood, a brief description of the operation of the entire circuit from the microphone input to the telephone line or transmitter will be given.

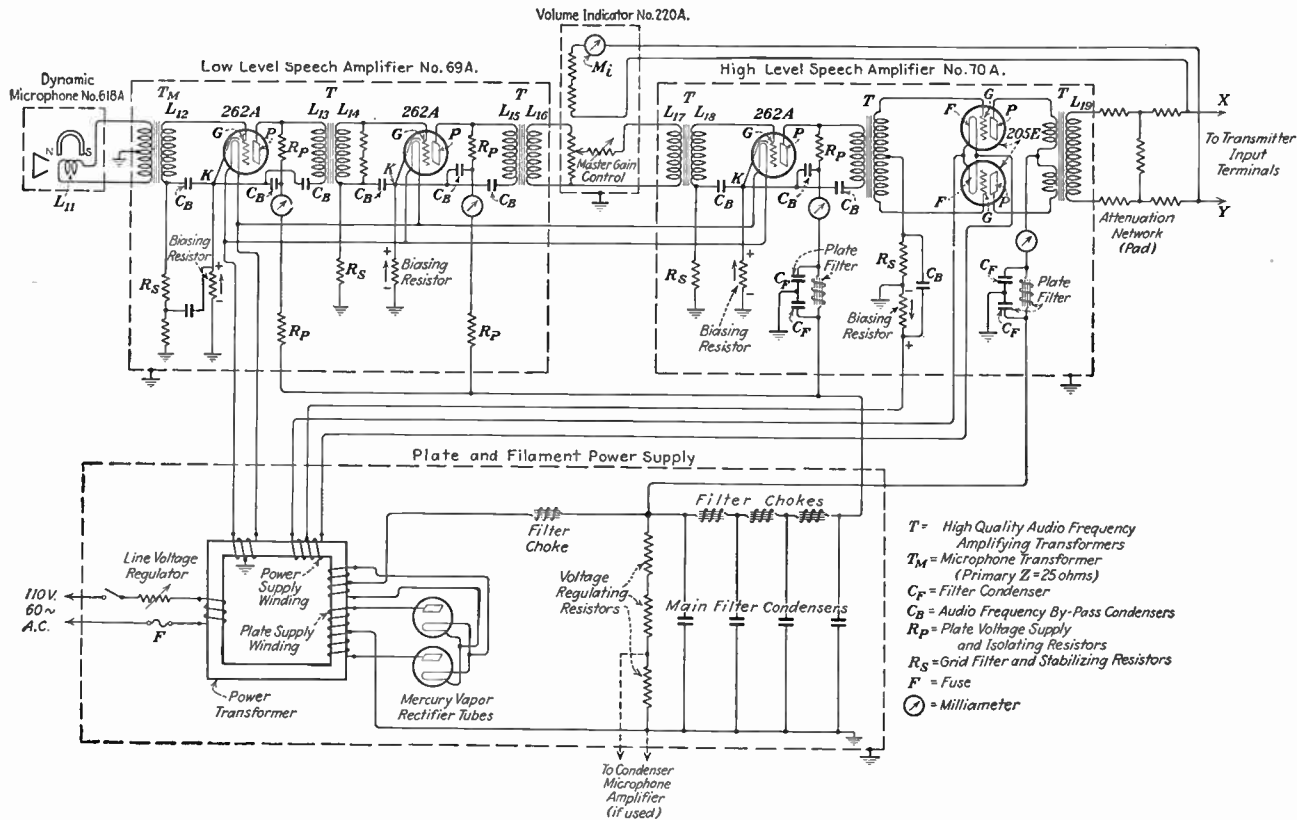


FIG. 247.—Simplified diagram of Western Electric speech amplifiers.

Assuming that the primary winding of the power transformer is energized by closing the alternating-current supply-line switch, and that as a result the proper filament and plate voltages are developed across the respective secondary windings, then, through the rectifying action of the mercury-vapor tubes, and the filtering action of the choke coils and condensers, steady plate voltages will be applied between the cathodes and plates of all tubes. This causes a steady current to flow in all plate circuits, and also develops a grid bias on all tubes through the current flow in the respective resistances connected in the cathode-return circuits of the tubes.

With this picture in mind let us see what happens during operation. When sound waves are impressed upon the diaphragm of the dynamic microphone, the small coil in the unit is moved to and fro at a rate depending upon the frequency of the impressed sound waves. Now since this coil is correctly situated in the magnetic field of a permanent magnet, the to-and-fro motion of the coil in the field will induce an alternating e.m.f. into it in accordance with the laws of induction. This action results in the flow of an alternating current in the primary winding of the microphone input transformer, T_M , and creates a moving magnetic field about it. This moving field induces an alternating e.m.f. across the secondary winding, L_{12} . Now since one end of the secondary winding is connected to the grid of the first tube, and the other end of the winding is connected to the cathode through the grid-filtering and by-pass condenser, C_B , the alternating voltages will be applied across G and K of the first tube. This voltage will develop an alternating variation in the plate circuit of this tube and the associated primary winding, L_{13} , of the transformer. Now since this primary winding is connected in a parallel-feed arrangement, that is, with a condenser in series with the winding, only the alternating or signal voltage will pass through it. This arrangement, therefore, isolates the normal plate-current flow from the primary winding owing to the blocking action of the condenser, C_B , to a direct current, and thereby prevents saturation of the iron core of the transformer. This arrangement is very necessary when high-grade magnetic cores are used, since without such a circuit the core would saturate very readily, and produce serious distortion owing to the effective decrease of the inductance of the transformer. The normal plate-current flow, therefore, takes place only through the feed and plate-filtering resistors, R_P , as shown.

The signal variations now present in the primary winding are induced into the secondary winding, L_{14} , and applied across the G and K elements of the second tube. These increased variations now present in the plate circuit of the second tube are then passed through the associated primary winding, L_{15} , and series condenser, C_B , as before. These variations are then transferred to the secondary winding, L_{16} , and a resultant

voltage is developed across this winding. However, this voltage is not directly applied to G and K of the next (third) amplifier tube, as before, but is sent through an attenuation system and another transformer. This attenuation system is generally referred to as the master gain control, since it allows the proper signal level to be applied to the grid of the first tube in the higher-level amplifier (third tube). This master-gain-control potentiometer varies the input in steps of 2 db each, and can be varied over a range of 18 steps, or 36 db. The proper adjustment of this control is determined by the level required, and is indicated by the volume indicator meter, M_i , as will be seen later.

The variations now present in the primary winding, L_{17} , of the input transformer to the third tube are then induced across the secondary winding, L_{18} , and applied to G and K , also as before. The greatly increased voltage variations now present across G and K of this tube produce variations through the plate and its associated parallel-fed primary circuit. These variations are then induced into the secondary winding of the push-pull input transformer. Here they are alternately fed to the grids of two push-pull tubes, 180 degrees out of phase, as explained in Sec. 144. The resultant plate variations in the plate circuit of the push-pull tubes flow in their associated primary winding, and are then transferred by induction to the secondary winding, L_{19} . The signal-voltage variations now developed at the output terminals of the high-level amplifier are attenuated through a 5-db attenuation network. Here they are simultaneously applied to the line and volume indicator. These alternating variations flowing into the indicator circuit are rectified by a small copper oxide type of rectifier so that the current flowing through the moving coil of the instrument is of a direct pulsating nature.

A level key is provided with the instrument so that various resistance combinations may be used to obtain proper deflections for different power levels. When shunted across a 500-ohm line, or an amplifier which has an output impedance of 500 ohms, power-level indications of -8 , 0 , and $+10$ db may be obtained. The scale is calibrated over the upper portion which reads, from left to right -2 db, -1 db, 0 db, and $+2$ db. Hence, if the volume-control potentiometer between the low-level and high-level amplifier is varied so that the needle of the volume indicator causes a deflection of -1 db at peak program pulses, then, if the level key is in the $+10$ db position, the energy level is $+9$ db. This energy level may readily be converted into a power expression as previously shown in Sec. 192. Thus since zero level is generally referred to as 0.006 watt or 6 mw., then a $+9$ db level may be expressed as a power level of approximately 0.05 watt or 50 mw.

The lower-level section of the amplifier is designed for an input and output impedance of 200 ohms, and has a gain of approximately 60 db. The high-level section of the amplifier, on the other hand, is designed

for an input impedance of 200 ohms, and an output impedance of 500 ohms.

A symmetrical wave shape practically devoid of distortion up to a power level of 240 mw. or +16 db may be obtained with this amplifier with a maximum gain of approximately 50 db.

247. The Frequency Monitor.—Figure 248 illustrates the wiring diagram of the W.E. type 1-A frequency-monitoring unit used in conjunction with the 12-A transmitter. The monitor employs a calibrated crystal and temperature-control chamber identical with that explained in Chap. IV on Transmitting Circuit Principles, which is interchangeable with the crystal oscillator of the 12-A transmitter unit. In addition to the oscillator, an audible and visual indicator is provided to show the frequency variation of the carrier wave, two screen-grid radio-frequency amplifiers, one detector, one argon regulator, and one mercury-vapor full-wave power-rectifier tube. The entire unit is operated from an alternating-current, 110-volt or 220-volt, 60-cycle power-supply system.

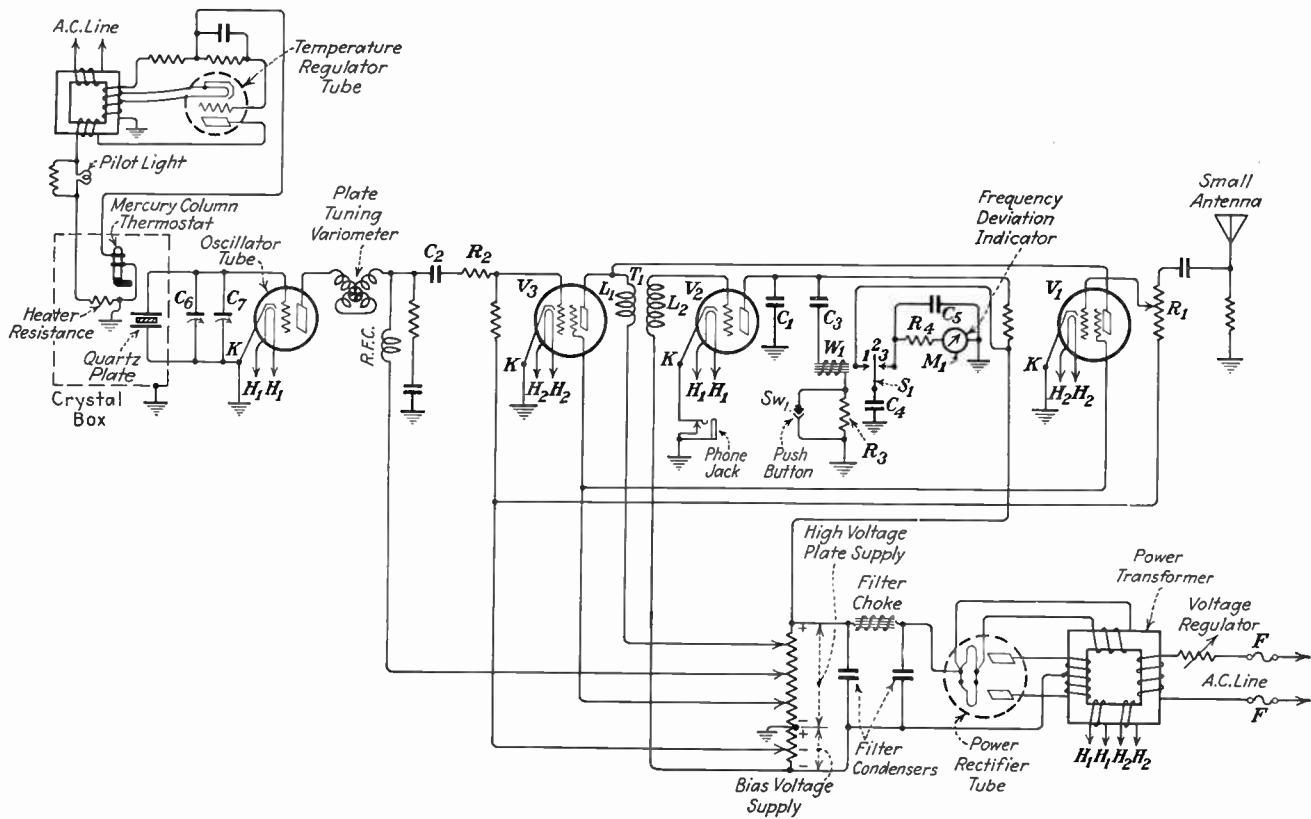
The monitor unit is capable of indicating a frequency deviation of ± 75 cycles of the carrier wave from its assigned frequency.

A small antenna is generally used to pick up the radio-frequency voltages from the transmitter, although it is also possible to determine variations by connecting the monitor to a receiver tuned to the transmitter frequency.

The theoretical operation of the monitor unit is as follows. When the radio-frequency voltages from the carrier frequency are picked up by the small antenna, they are applied across the grid-cathode terminals of the screen-grid tube V_1 , the intensity of the voltage being regulated by the potentiometer R_1 . These voltage changes then produce plate variations at a radio frequency in the tube which pass through the primary winding of the radio-frequency transformer T_1 . The voltages are then applied across the grid and cathode of tube V_2 by induction from L_1 to L_2 . This tube is so biased negatively that the radio-frequency voltages on the grid produce asymmetrical pulses in the plate circuit (detection).

Now let us assume that these variations are the result of a carrier frequency of 500,000 cycles. Although this frequency is rectified into unilateral pulses, the sequence of the positive peaks is of such rapidity as to be beyond audibility or visibility. Hence, these high-frequency pulses cannot pass through the relay winding W_1 , owing to its high reactance to high frequencies, but will pass through the radio-frequency plate by-pass condenser C_1 .

Assuming that the crystal oscillator is in operation at the same frequency, 500,000 cycles, the high-frequency voltage developed in the output of the oscillator will be applied across the grid-cathode terminals of the radio-frequency amplifier tube V_3 through the coupling condenser



C_2 and the feed resistance R_2 . This will cause another frequency variation of 500,000 cycles to be present in the plate circuit of V_2 . This frequency too will induce an e.m.f. across the detector tube V_2 as before. Thus two sustained frequencies, namely, the one from the transmitter and the one from the oscillator, are present in the plate circuit of the tube V_2 . Hence, since both of these frequencies are assumed to be identical, their resultant rectified plate variations will pass through the by-pass condenser C_1 .

Let it now be assumed that the frequency of the transmitter is 25 cycles higher than that of the oscillator standard, or 500,025 cycles. Thus two radio-frequency voltages of 500,000 and 500,025 cycles will be developed across the grid and cathode of detector tube V_2 with their resultant variations flowing in the plate circuit through C_1 . However, since both of these frequencies are flowing through C_1 , they beat against one another and develop across the condenser periodic additions and cancellations of voltage to the value of the difference between the two radio frequencies. The resultant difference between the two frequencies is in the low-frequency range, that is, 25 cycles. Hence, the low-frequency rectified pulses will readily pass through the reactive circuit C_3 , W_1 , and R_3 and energize the sensitive relay S_1 . (R_3 must be short-circuited by closing frequency-test switch Sw_1 to pass sufficient current to energize W_1). This action will attract the vibrating arm and close the contacts 1 and 2 and charge the condenser C_4 . At the instant the low-frequency pulse through W_1 ceases, the arm returns and closes contacts 2 and 3, allowing the condenser C_4 to discharge through the frequency indicator M_1 . Hence the periodic low-frequency variations of the rectified pulses through W_1 will allow the condenser C_4 to discharge at a definite rate and in one direction only through the direct-current meter M_1 . The condenser C_5 and the resistance R_4 control the discharge rate of the charging condenser C_4 to produce the proper damping effect on the moving element of the meter M_1 .

Now, if the frequency of the carrier wave is 499,975 cycles, or 25 cycles below the frequency-monitor standard of 500,000 cycles, the meter M_1 will still indicate the rectified pulses of 25 cycles. If, therefore, it is desired to ascertain if the frequency variation of the transmitter is -25 or $+25$ cycles the frequency of the monitor crystal must be slightly altered. This is done by inserting an extremely small capacity C_6 across the crystal. This condenser is operated by the same push button which closes the test switch Sw_1 . When the push-button is gently pressed the small capacity C_6 is varied toward minimum capacity by separating two small circular metal plates. When the push-button is released, the capacity of C_6 increases very slightly as the plates move closer together.

Therefore, if the frequency-deviation indicator switch Sw_1 is closed and the frequency indicator M_1 reads 25 cycles, the push-button is very

slightly depressed and released, and the deflection of the indicator is again noted. If it tends to move slightly upward, it is an indication that the transmitter frequency is 25 cycles above that of the monitor standard, or +25 cycles. The reason for this action may be clearly seen if we recall that the crystal represents the equivalent of a tuned circuit in which the frequency is dependent upon the constants L and C . For example, if the value of C across an inductance is increased, the frequency period of the circuit will decrease. Hence, the small capacity C_6 across the crystal very slightly alters the monitor frequency and enables the operator to determine the direction of the frequency shift.

A small trimming condenser C_7 is connected across the quartz crystal. This condenser is permanently adjusted to half-capacity at the time of frequency calibration so as to provide flexibility, if it is found necessary to increase or decrease very slightly the oscillator frequency.

If desired, the monitor frequency may be checked periodically to insure that its indication gives a true reading of the transmitter frequency at all times. Checking service may be had for a fee from several sources among which is the RCA Communications at Riverhead, L.I., N.Y. or at San Francisco, Calif.

If an immediate check on the exact transmitter frequency is made by telephone, the crystal monitor may be immediately adjusted against the transmitter frequency owing to the fact that the transmitter is effectively standardized at that moment. The condenser C_7 on the monitor is adjusted until the monitor-frequency indicator reads exactly the same as that reported by the standard measurements. After this has been done, the monitor is calibrated with the standard and the transmitter may then be adjusted to minimum frequency variation, within ± 10 cycles as now indicated by the crystal monitor.

248. Modulation Measurements.—When the entire transmitter has been properly adjusted to the desired carrier-frequency stability and power requirements, a careful check of the modulation percentage must be made.

There are several practical methods of determining approximately if the transmitter here described is modulating over 85 per cent and not more than 100 per cent.

The simplest method and a fairly accurate one is that in which a single-tone frequency such as is supplied by the 60-cycle power line is fed into the radio-frequency stage to be modulated. Assuming that before modulation the output mesh current in the amplifier circuit is 1 amp. (this would be the exact amount of current necessary to indicate that 100 ohms are present in this circuit to dissipate the proper amount of power, namely, I^2R or 100 watts, and to excite the grids of the power-amplifier tubes) then, if the 60-cycle single tone is supplied directly to these tubes through the medium of the grid-modulation transformer

T_{14} , and if the mesh current in the plate circuit, as indicated by the meter M_2 , reads 1.225 times (22.5 per cent rise) the normal amount of 1 amp., it is a fair indication that the carrier frequency is modulated 100 per cent. Hence, if, when the single tone is applied, the mesh or tank current rises to 1.225 amp., the transmitter is most certainly modulated at or very near to 100 per cent. Perhaps an even better check when using this method is to use the radio-frequency ammeter M_3 in the plate circuit of the power-amplifier tubes. For example, if this meter should read 3.2 amp. during the period in which the carrier frequency is unmodulated, then, if a 60-cycle single tone is supplied (by increasing the gain of the potentiometer control in the speech-input circuit) until the meter rises to a value 22.5 per cent higher than 3.2, or 3.9, amp., the carrier is modulated at approximately 100 per cent.

It is apparent that this method would be practically useless if anything other than a pure single tone were used owing to the difficulty of the meter to follow the peaks during complex modulation.

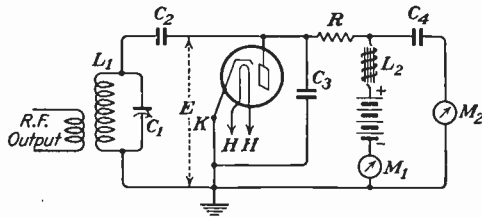


FIG. 249.—Vacuum-tube peak voltmeter for indicating percentage of modulation.

Various types of indicators called *peak voltmeters* or *modulation meters* have been designed to determine accurately the percentage of modulation by this method. Some of these instruments enable a constant check on the modulated frequency, even during the complex modulation while the program is in operation. This is accomplished by a time-lag circuit of capacity and resistance in conjunction with the indicating meter so that, when the needle follows a peak, it will momentarily stay in its deflected position and thereby permit a fair check on the modulation percentage during peaks.

Figure 249 illustrates the wiring diagram of a vacuum-tube peak voltmeter for indicating the percentage of modulation.

The r.-f. output of the final stage of the transmitter is magnetically coupled to the input circuit of the voltmeter so that, when L_1 and C_1 are adjusted to the carrier frequency, a potential will be developed across the parallel mesh. This e.m.f. will be applied across the plate-cathode terminals of the rectifier tube through the medium of the direct-current-blocking, radio-frequency, by-pass condenser C_2 . The steady direct current flowing in the plate circuit through L_2 and R establishes a steady potential across them. If the radio-frequency voltage is unmodulated,

no voltage variations will be developed across the resistance R owing to its high resistance and associated impedance L_2 . Hence, these radio-frequency variations will pass through the relatively low-reactance path of C_3 .

If, however, the radio-frequency carrier is modulated, the current through the resistance R will be varied in direct proportion to the audible rate at which the carrier is varied. Since this results in series of low-frequency pulses across C_4 and the meter M_2 , the condenser C_4 will discharge these pulses at an audio-frequency rate through the meter. Thus, if a single tone is used to modulate the carrier frequency, a direct scale reading of the effective modulation may be obtained by taking the ratio of the current through the direct-current meter M_1 to that of the alternating-current meter M_2 and multiplying it by the product of $\sqrt{2} \times 100$, thus,

$$\text{Modulation} = \left(\sqrt{2} \times \frac{I_2}{I_1} \times 100 \right) \text{ per cent} \quad (173A)$$

The most accurate method, however, for determining the modulation percentage of a radio transmitter is by the use of the cathode-ray oscil-

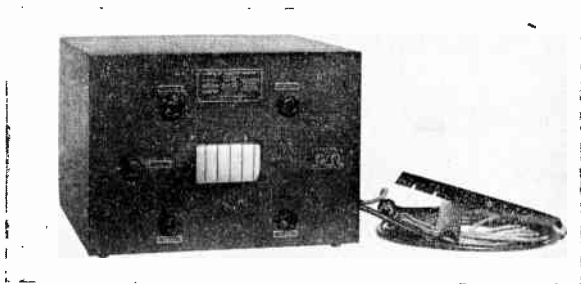


FIG. 250.—RCA Victor cathode-ray modulation indicator. (Courtesy of RCA Manufacturing Company, Inc.)

lograph. With this device an absolutely accurate indication of modulation percentage may be obtained at all modulated frequencies since the cathode ray operates at radio-frequencies and is not restricted to the sluggish audio frequencies. In other words, the necessity of rectification and its resultant damping losses are entirely obliterated because the indication is produced by the radio-frequency carrier frequency itself.

Figure 250 illustrates the RCA cathode-ray modulation indicator type 49-A which utilizes an RCA-904 cathode-ray tube. In this indicator a sweep circuit is employed using electromagnetic deflection as shown in Fig. 251. The modulation percentage indication is normally obtained by the operation of the vertical sweep connected to a 115-volt 60-cycle power supply, but it may be connected to some other source, if desired.

The radio-frequency carrier is picked up by a tuned circuit designed to cover the entire broadcast and police bands through the medium of a tapped coil. Three voltage-regulator circuits and a sweep-circuit potentiometer are provided for the necessary adjustments. When these controls have been carefully regulated to the carrier frequency, a continuous wave form will appear on a fluorescent screen as shown in the illustration. This panel has three red lines and two black lines which serve to indicate the modulation percentage on both the positive and negative peaks of the cycle.

The unmodulated carrier frequency is adjusted so that it extends on both halves of the cycle to the black lines as illustrated in Fig. 252(a). Now let us assume that a single-tone frequency of 60 cycles is applied through the speech-input equipment with the proper output level to

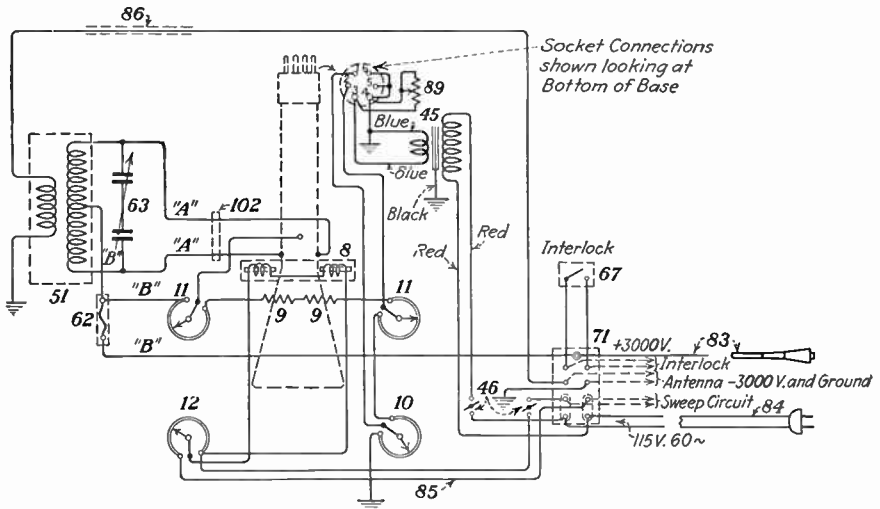


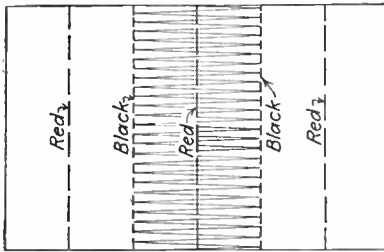
Fig. 251.—Sweep circuit of RCA Victor cathode-ray modulation indicator (Type 49-A).
(Courtesy of RCA Manufacturing Company, Inc.)

modulate the carrier. If the carrier frequency is varied in amplitude so that the peaks extend to the outer red lines, the carrier is said to be positively modulated at 100 per cent. If, at the same time, the decreased peaks or troughs extend to the center red line, then 100 per cent negative modulation is obtained. In other words, if the transmitter carrier frequency is completely modulated at 100 per cent, both the positive and negative variations must extend to the outer and center red lines, respectively. This condition is illustrated in (b).

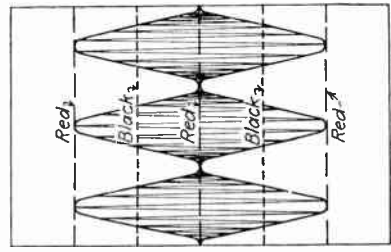
If, on the other hand, the carrier frequency is modulated only 50 per cent, the positive and negative peaks will extend only to half the value between the black and red lines as indicated in drawing (c). If the modulation is in excess of 100 per cent, the positive peaks will extend

beyond the outer red lines and the negative peaks will be entirely ineffective as illustrated in drawing (d). This last condition effectively increases the damping of the modulated wave and introduces serious distortion.

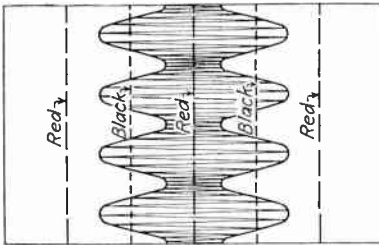
Precautions: If the beam is allowed to play continuously on one spot of the fluorescent screen, the active coating may be permanently damaged. The position of the radio-frequency coupling coil, item 51, should not be adjusted without removing the high voltage as it operates at a potential of 3,000 volts direct current.



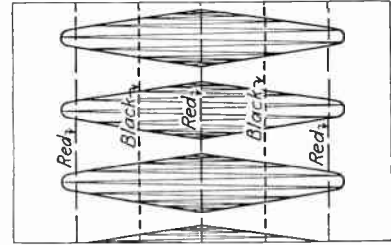
(a)-Unmodulated Carrier



(b)-Carrier Modulated 100 Per Cent



(c)-Carrier Modulated 50 Per Cent



(d)-Carrier Modulated Over 100 Per Cent

FIG. 252.—How carrier wave and modulation appears on fluorescent screen of RCA Victor cathode-ray modulation indicator. (Courtesy of RCA Manufacturing Company, Inc.)

249. RCA Victor Broadcast Transmitters. *a. Type 1-D, 1,000 Watts.*—This transmitter consists of two self-contained units, a 100-watt transmitter unit (ET-4240) and a 1,000-watt amplifier unit (AA-4251). By substituting the proper tubes, the transmitter unit can be used alone as a 250-watt transmitter (ET-4250). When so used its output can be decreased to 100 watts by means of a switch located on the control panel. This is an advantageous feature when the station is authorized to use a larger power output during the daytime than at night.

A complete schematic circuit diagram of the 1-kw. transmitter complete is shown in Fig. 253. Most of the significant features of the transmitter will be evident from a study of these diagrams, and it is not the intention here to go into long descriptions of characteristics which are obvious from explanations previously given.

This transmitter is designed to modulate 100 per cent with an audio input of approximately 0-db level (0.006 watt).

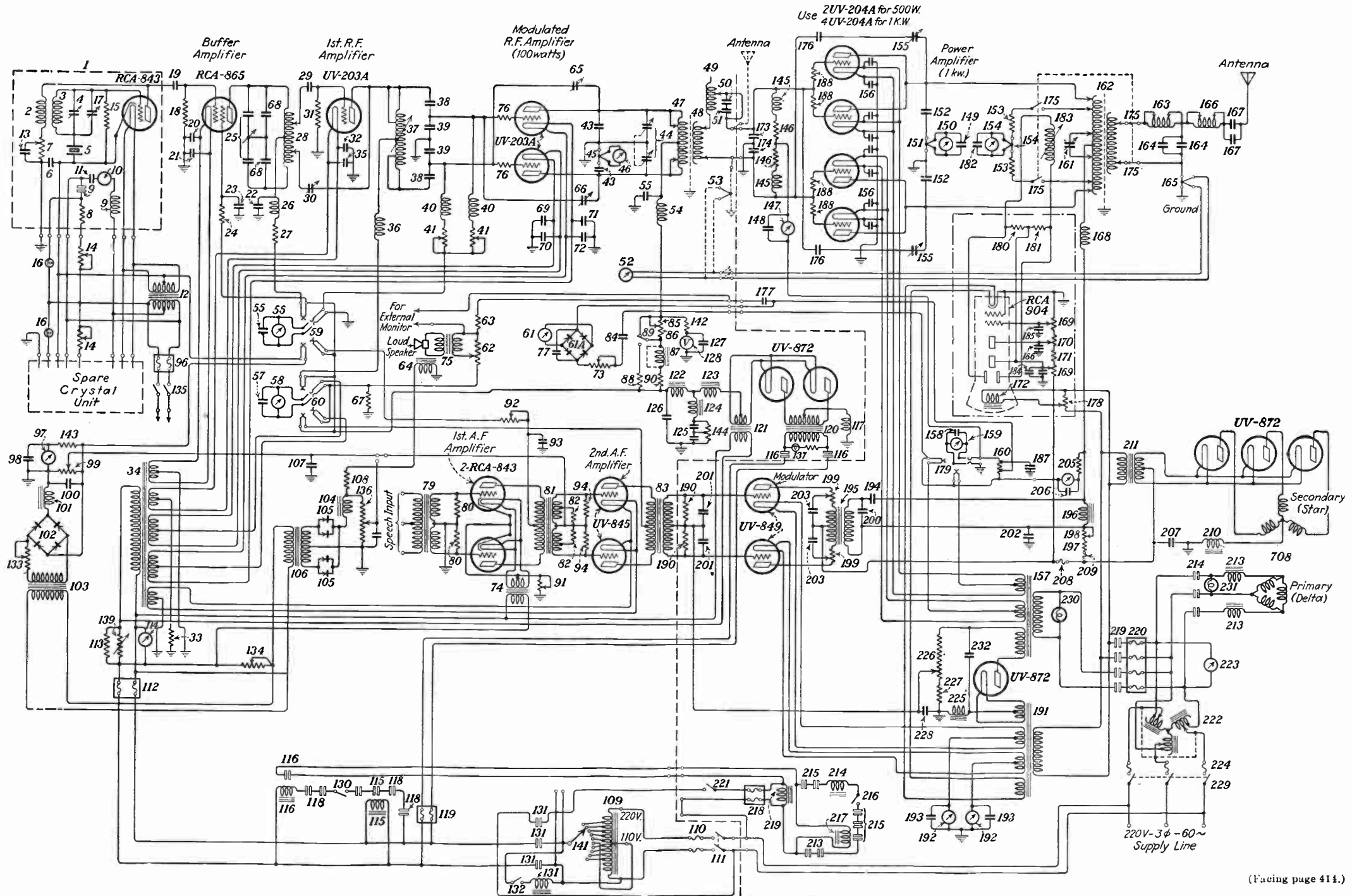


FIG. 253.—RCA Victor 1-kw. broadcast transmitter. (Courtesy of RCA Manufacturing Company, Inc.)

(Facing page 414.)

PARTS OF RCA VICTOR BROADCAST TRANSMITTER (See Fig. 253)

1	Crystal oscillator unit.	58	D-c. milliammeter, 0 to 500 ma.	116	Plate contactor.	176	Neutralizing condenser, 0.0001 mf.
2	Oscillator plate inductance.	59-60	Meter switch.	117	D-c. overload relay, dash-pot type.	177	Monitor blocking condenser, 28 mf., 1,000-volt d-c. rating.
3	Oscillator grid inductance.	61	Modulation indication meter, 0 to 1.5 ma.	118	Door switches.	178	Potentiometer, 2,300 ohms.
4	Oscillator tank circuit capacity, 0.000365 mf.	61A	Modulation indication meter Rectox rectifier units.	119	Cartridge-type fuses, 20 amp., 250 volts.	179	Meter switch.
5	Quartz plate.	62	Loud-speaker volume-control potentiometer, 500 ohms, 50 watts.	120	Plate-supply power transformers.	180-181	Coupling-coil balancing resistance.
6	By-pass condenser, 0.01 mf., 2,000-volt rating.	63	Monitor resistances, 30 ohms, 20 watts.	121	Filament transformer.	182	By-pass condenser, 0.01 mf.
7	Plate feed-back resistance, 1,000 ohm, 3 watts.	64	Monitoring loud-speaker.	122-124	Filter choke coil.	183	Coupling coil for cathode-ray modulator indicator.
8	Crystal heater resistances, 500 ohms.	65-66	Neutralizing condenser, 0.000056 mf.	125	Filter condenser, 2 mf., 1,500-volt d-c. rating.	184	By-pass condenser, 0.001 mf., 5,000-volt rating.
9	Heater control relay.	67	Bleeder resistance, 25 ohms, 25 watts.	126	Filter condenser, 10 mf., 1,000-volt d-c. rating.	185	By-pass condenser, 0.01 mf., 2,000-volt rating.
10	Mercury thermo-regulator.	68	Condenser for buffer amplifier tank.	127	By-pass condenser, 0.01 mf.	186	By-pass condenser, 0.01 mf., 2,000-volt rating.
11	Relay condenser, 0.2 mf., 300-volt rating.	69-72	By-pass condenser, 0.01 mf., 2,000-volt rating.	128	Plate voltmeter, 0 to 1,500 volts.	187	Power-supply filter condenser, 10 mf., 1,000-volt rating.
12	Filament transformer, 50 or 60 cycles.	73	Modulation indicator adjustable resistance, 10,000 ohms, 75 watts.	129	Electrolytic condenser, 200 mf., 50-volt rating.	188	Power-amplifier grid resistance, 100 ohms, 5 watts.
13	By-pass condenser, 0.01 mf., 2,000-volt rating.	74	Filament transformer.	130	Plate-voltage switch.	189	Modulator (not separately numbered on diagram).
14	Heater control resistance, 500 ohms, 60 watts.	75	Loud-speaker input transformer.	131	Filament contactor.	190	Secondary stabilizing resistances.
15	Oscillator grid leak resistance, 1 meg-ohm, 1 watt.	76	Grid resistance, 100 ohms, 5 watts.	132	Transmitter starting switch.	191	Filament transformer.
16	Heater indicator lamp.	77	By-pass condenser, 0.01 mf., 2,000-volt rating.	133	Rectox regulator resistance.	192	Plate ammeter, 0 to 300 ma.
17	Frequency vernier capacity, 0.00007 mf.	78	A.F. amplifier and modulator (not separately numbered on diagram).	134	Regulator resistance for T103 and T106.	193	By-pass condenser, 0.01 mf.
18	Buffer amplifier grid resistance, 100,000 ohms, 1 watt.	79	Audio-input transformer.	135	Crystal-heater switch.	194	Blocking condenser, 6 mf., 2,000-volt rating.
19	Coupling capacity, 0.00004 mf., 5,000-volt rating.	80	Audio-input load resistance, 9,000 ohms, 1 watt.	136	Biasing potentiometer, 5,000 ohms, 50 watts.	195	Modulation output transformer.
20-23	By-pass condenser, 0.01 mf., 2,000-volt rating.	81	Interstage coupling transformer.	137	Plate-voltage indicator light.	196	Modulation reactor.
24	Screen grid resistance, buffer amplifier, 50,000 ohms, 44 watts.	82	Load resistance, 15,000 ohms, 3 watts.	138	Spare-crystal indicator light.	197, 198	Plate-feed resistances.
25	Split stator tank circuit capacity, 0.00048 mf. each section.	83	Modulator interstage coupling transformer.	139	Filament rheostat, 10 ohms, 75 watts. Adjustable.	199	Modulation plate resistance, 100 ohms, 30 watts.
26	Plate choke coil.	84	Blocking condenser, 2 mf., 2000 volt rating.	140	Meter panel lights.	200	Secondary shunt condenser.
27	Plate resistance, 15,000 ohms, 100 watts.	85, 86	Amplifier feed resistance.	141	Line-voltage switch.	201, 202	Secondary by-pass condenser.
28	Plate inductance.	87	Modulator reactor.	142	Voltmeter multiplier for 0 to 1,500-volt meter.	203	Primary by-pass condenser.
29	Blocking condenser, 0.001 mf., 5,000-volt rating.	88	Shunt reactor resistance.	143	Voltmeter multiplier for 0 to 500-volt meter.	204	Power equipment (not separately numbered on diagram).
30	Neutralizing condenser, 0.000034 mf.	89	Power changing switch.	144	Filter resistance, 500,000 ohms, 3 watts.	205	Voltmeter and multiplier, 0 to 5,000 volts, d.c.
31	Grid resistance, 2,000 ohms, 5 watts.	90	Power-amplifier plate resistance, 2,500 ohms with 10 taps, 200 watts.	145	R.F. choke coil.	206	By-pass condenser, 0.01 mf.
32	By-pass condenser, 0.01 mf., 2,000-volt rating.	91	Biasing resistance, 750 ohms, 100 watts. Variable.	146	Grid resistance, 2,000 ohms, 60 watts.	207	Filter condenser, two 3.5-mf. units, 3,000-volt rating.
33	Biasing resistance, 100 ohms, 5 watts.	92	Plate feed resistance.	147	D.c. ammeter, 0 to 300 ma.	208, 209	Fuse, 1 amp., 2,000 volts.
34	Filament transformer.	93	By-pass condenser, 4 mf., 1,000-volt d-c. rating.	148	By-pass condenser, .01 mf.	210	Filter choke coil.
35	By-pass condenser, 0.01 mf., 2,000 volt rating.	94	Grid resistance, 100,000 ohms, 3 watts.	149	R.F. ammeter, 0 to 8 amp.	211	Filament transformer.
36	R.F. choke coil.	95	Power equipment (not separately numbered on diagram).	150	External thermocouple for 0- to 8-amp. meter.	212	Plate power transformers, 3.75-kva. rating.
37	Plate inductance.	96	Cartridge-type fuses, 6 amp., 250 volts.	151	Power-amplifier tank circuit condenser.	213	Overload relays, oil dashpot, time-delay relay.
38	Coupling condenser.	97	D-c. voltmeter, 0 to 500 volts.	152	Phantom antenna resistance, twelve 220-ohm units in series.	214	Plate contactor.
39	By-pass condenser.	98	Meter by-pass condenser, 0.01 mf.	153	Phantom antenna ammeter, 0 to 1 amp.	215	Door interlocking switches.
40	R.F. choke coil.	99	Potentiometer, 3,500 ohms, 50 watts.	154	Variable condenser, 0.00011 mf.	216	Plate on switch.
41	Power amplifier grid leak resistance, 20,000 ohms, 75 watts.	100	Filter condenser, 18 mf., 1,000-volt rating.	155	By-pass condenser, 0.01 mf., 2,000-volt rating.	217	Time-delay relay, oil-dashpot type.
42	Spare crystal unit. Same as 5.	101	Filter choke coil.	156	Filament and bias rectifier transformer	218	Fuses, 6 amp., 250 volts.
43	Tank circuit by-pass condenser.	102	Rectox rectifier unit.	157	By-pass condenser, 0.01 mf.	219	Filament contactor.
44	Tank circuit tuning condenser, 0.00022 mf.	103	Rectifier transformer.	158	Power-amplifier plate ammeter, 0 to 1 amp.	220	Fuses, 6 amp., 250 volt.
45	Thermocouple, 0 to 5 amp.	104	Filter choke coil.	159	Biasing resistance, 150 ohms, 200 watts.	221	Filament switch.
46	R.F. ammeter, 0 to 5 amp.	105	Rectox rectifier units.	160	Variable tank-circuit capacity, 0.00022 mf.	222	Autotransformer.
47	R.F. transformer.	106	Rectifier transformer.	161	R.F. transformer.	223	Line voltmeter, 0 to 250 volts a.c.
48	Antenna coupling coil.	107	Filter condenser, electrolytic type.	162	Harmonic suppression inductance.	224	Fuses, 20 amp., 250 volts.
49	Loading inductance.	108	200 mf., 50-volt rating.	163	Antenna-coupling condenser.	225	Bias filter choke coil.
50-51	Antenna series capacity, 0.001 mf.	109	Voltage divider resistance section, 40 ohms, 100 watts.	164	Antenna ammeter thermocouple.	226	Bias potentiometer, 125 ohms.
52	Antenna radio-frequency ammeter.	110	Autotransformer.	165	Antenna loading inductance.	227	Bias resistance, 200 ohms, 200 watts.
53	Thermocouple for antenna r.f. ammeter (not shown).	111	Cartridge-type fuses, 30 amp., 250 volts.	166	Antenna condenser.	228	Bias filter condenser, 200 mf., 200 volts, electrolytic type.
54	R.F. choke coil.	112	Line switch.	167	R.F. choke coil.	229	Line switch, three-pole single-throw variety.
55	Meter by-pass condenser, 0.01 mf.	113	Cartridge-type fuses, 10 amp., 250 volts.	168	Potentiometer, 100,000 ohms.	230, 231	Indicator lamp.
56	D-c. milliammeter, 0 to 150 ma.	114	Section of regulating resistance, 10 ohms, 100 watts.	169	Potentiometer, 350,000 ohms.	232	Bias filter condenser.
57	Meter by-pass condenser, 0.01 mf.	115	Filament voltmeter. 150 volts, A.c.	170	Separating resistance.	233	Panel lights.
			Time-delay relay. Dash-pot type.	171	Cathode-ray sweep-circuit coils		
				172	Grid tuning condenser.		
				173	Grid tuning condenser.		
				174	Phantom antenna switch.		
				175			

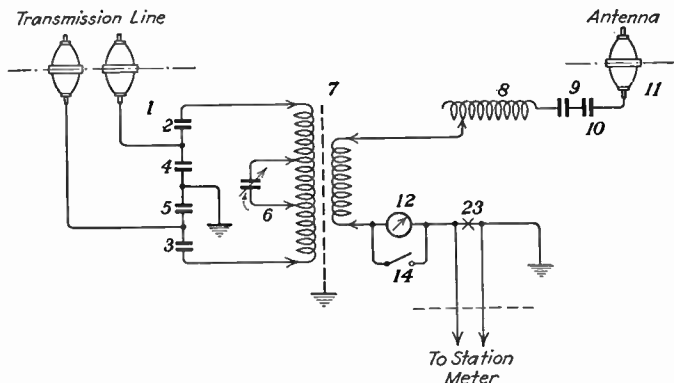


FIG. 254.—Transmission-line terminating equipment and antenna circuit. (Courtesy of RCA Manufacturing Company, Inc.)

PARTS OF T-L TERMINATING EQUIPMENT (SEE FIG. 254)

- 1 Insulators
- 2 } Condensers,
- 3 } value depends on
- 4 } frequency used
- 5 }
- 6 Tank tuning condenser, variable
- 7 Output r.-f. transformer
- 8 Tuning inductance
- 9 } Series antenna condensers
- 10 }
- 11 Insulator
- 12 R.-f. ammeter or thermogalvanometer¹
- 14 Switch for shorting out meter
- 23 Thermocouple

¹ This meter is used for tuning only and may be low range.

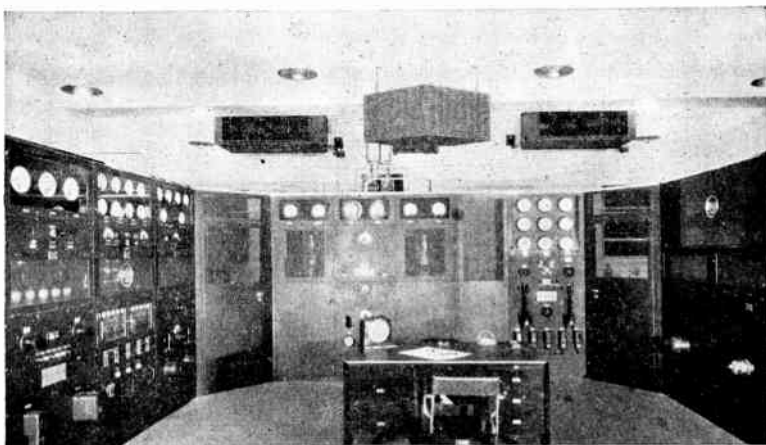


FIG. 255.—Typical arrangement of RCA Victor 50-kw. broadcast transmitter. (Courtesy of RCA Manufacturing Company, Inc.)

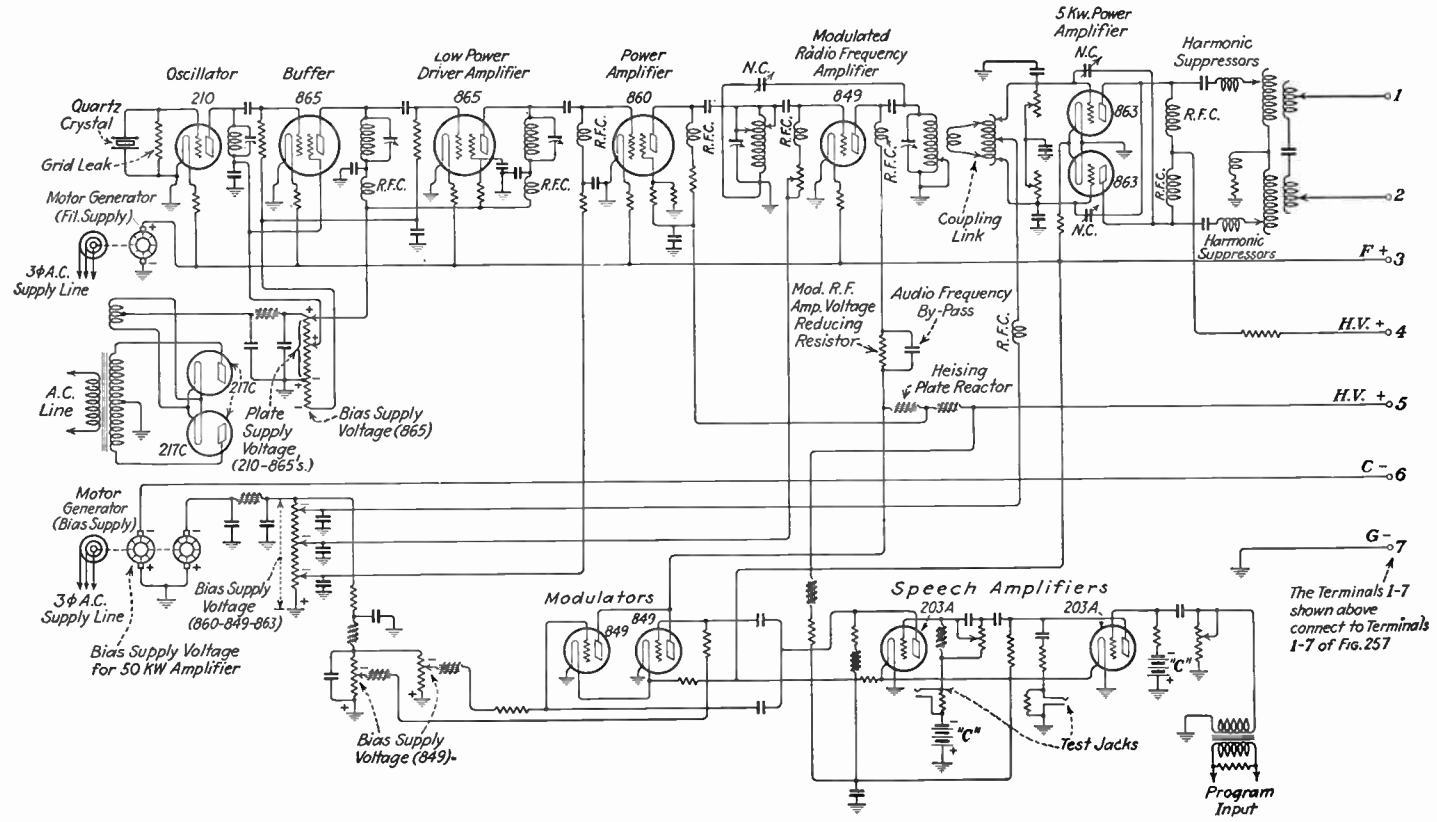


FIG. 256.—Simplified diagram of low-power section of RCA Victor 50-kw. broadcast transmitter.

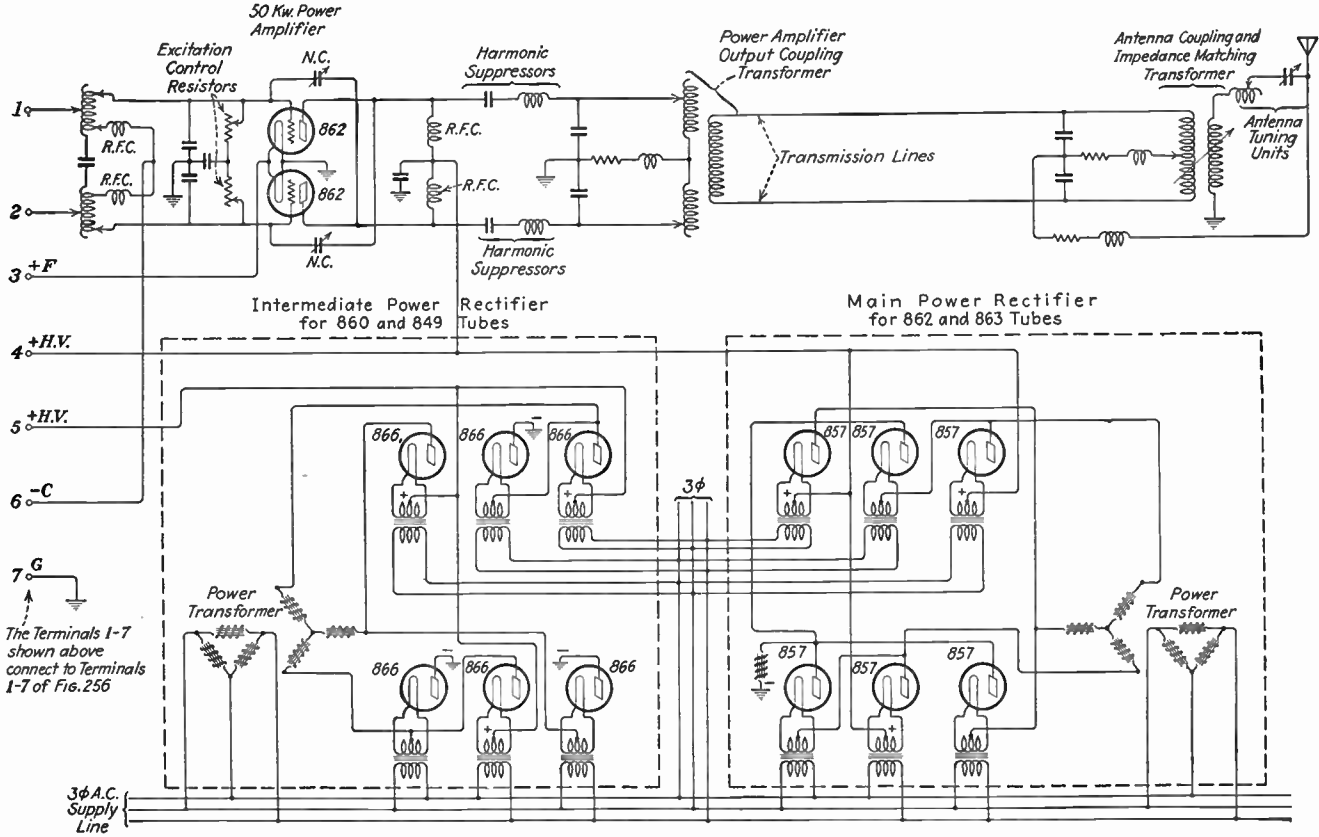


FIG. 257.—Simplified diagram of high-power section of RCA Victor 50-kw. broadcast transmitter.

The phantom antenna should be used to warm up the transmitter prior to broadcast periods, or when certain measurements or tests are to be made. This dummy load is connected to the output of the transmitter by operation of switch 175 which simultaneously disconnects the operating antenna. The phantom antenna consists of the items 153 and 154; see legend.

In Fig. 254 is shown a suitable circuit for transmission line or antenna connections to the output transformer of this transmitter.

OPERATING CHARACTERISTICS RCA BROADCAST TRANSMITTERS (SEE FIG. 253)

	Type ET-4240, 100 watts	Type ET-4250 250 watts	Type ET-4240 exciter unit
Line voltage.....	110 ± 5 volts	115 ± 5 volts	110 ± 5 volts
Oscillator plate current.....	7-15 ma.	7-15 ma.	7-15 ma.
Buffer amp. plate current.....	30-40 ma.	30-40 ma.	30-40 ma.
First r.-f. amp. plate current.....	75-125 ma	75-125 ma.	50-125 ma.
Second r.-f. amp. $\left\{ \begin{array}{l} \text{indirect measurement} \\ \text{of output} \dots\dots\dots \end{array} \right.$	200 ma.	417 ma.	150-250 ma.
total plate current $\left\{ \begin{array}{l} \text{direct measurement of} \\ \text{output} \dots\dots\dots \end{array} \right.$	150-225 ma.	350-425 ma	
Second r.-f. amp. total grid current.....	75-120 ma.	150-240 ma.	75-120 ma.
Second r.-f. amp tank current.....	.75-2.0 amp. r.f.	.75-2.0 amp. r.f.	.75-2.0 amp. r.f.
Second r.-f. amp. plate voltage.....	1,000 ± 20 volts	1,000 ± 20 volts	1,200 ± 50 volts
First audio amp. total plate current.....	90-100 ma.	90-100 ma.	40-50 ma.
Second audio amp. total plate current.....	40-50 ma.	40-50 ma.	115-120 ma.

	Type AA-4251 amplifier unit	
	1,000 watts output	500 watts output
Power amp. total grid current.....	175-300 ma.	100-160 ma.
Power amp. total plate $\left\{ \begin{array}{l} \text{indirect measurement of output} \\ \text{current} \dots\dots\dots \end{array} \right.$	835 ma.	417 ma.
$\left\{ \begin{array}{l} \text{direct measurement of output} \dots\dots\dots \end{array} \right.$	715-950 ma.	350-475 ma.
Power amp. tank current.....	2.70-3.65 amp. r.f.	2.70-3.65 amp. r.f.
Power amp. plate voltage.....	2,000 ± 20 volts	2,000 ± 20 volts
Modulation plate current (each tube).....	45-55 ma.	30-40 ma.

b. Fifty-kilowatt Transmitter, Model 50-B.—Any broadcast transmitter can be stepped up in power to 50 kw. or more by the addition of sufficient power amplification. A typical arrangement of the RCA Victor 50-kw. transmitter is shown in Fig. 255; the schematic diagram appears in Figs. 256 and 257.

The general theory of operation of this transmitter with regard to high- and low-frequency circuits is fundamentally the same as that of the lower power type previously explained. Only a general description, will here be given in view of the involved power supply, control and pro-

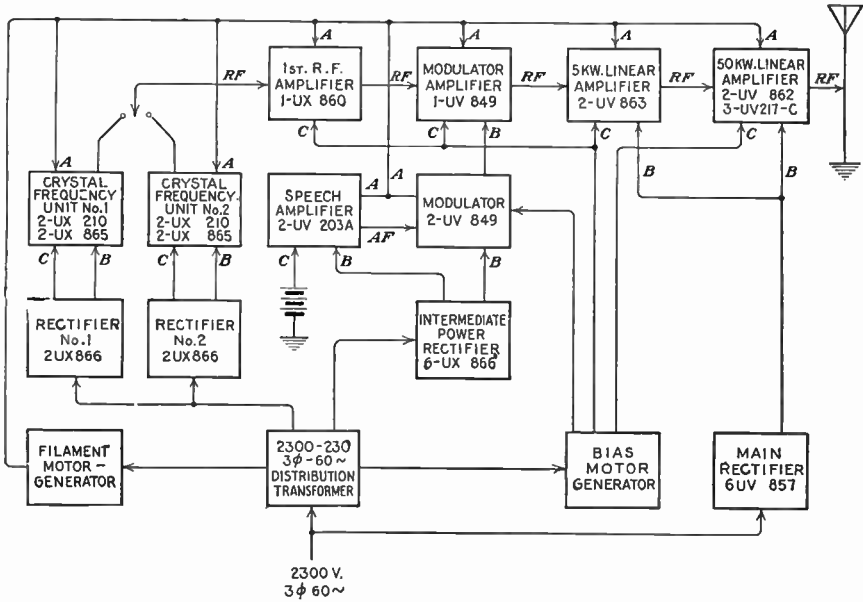


FIG. 258.—Box diagram showing component parts of RCA Victor 50-kw. broadcast transmitter. (Courtesy of RCA Manufacturing Company, Inc.)

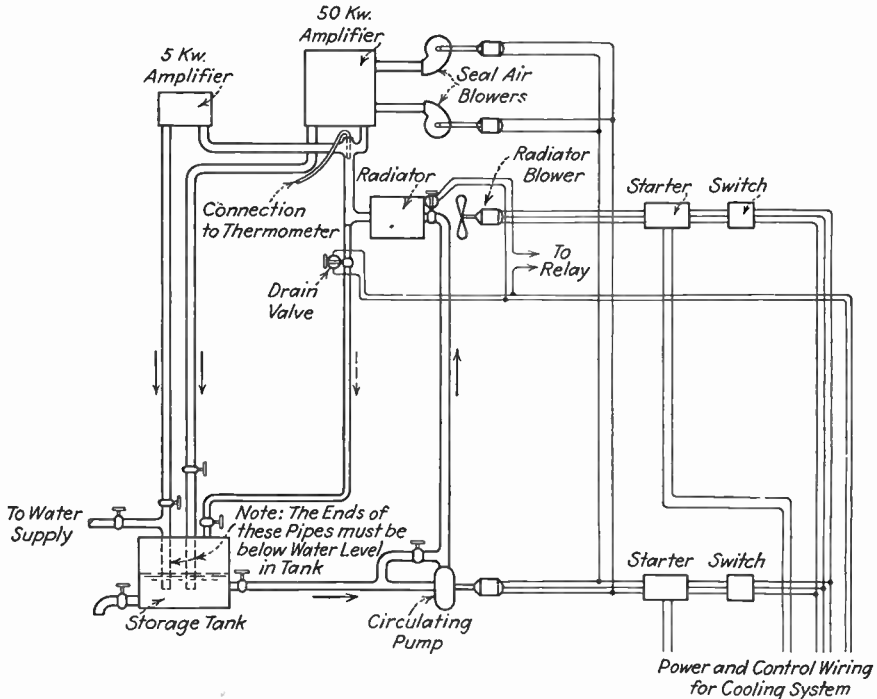


FIG. 259.—Water-cooling system of RCA Victor 50-kw. broadcast transmitter. (Courtesy of RCA Manufacturing Company, Inc.)

tective equipment used. Figure 258 illustrates a box diagram of the component parts and tubes used.

The power supply is divided into two branches: (1) 100-kva. circuit for the motor generator, control, and auxiliary power; (2) 258-kva. transformer for the main power rectifier. This power supply is controlled by two methods: automatic induction-voltage regulation and a step-starting resistance control. The voltages applied to the plates of the rectifier and amplifier tubes are applied gradually through a thermostatically controlled resistance unit which after an interval of time is short-circuited to permit the full value of supply-line voltage to be applied. During the interval in which this resistance is still in the circuit, the induction-voltage regulator is gradually increasing the voltage to the plates of the rectifier and power tubes, so that, when the resistance is finally short-circuited, the proper voltages will be applied to all plates.

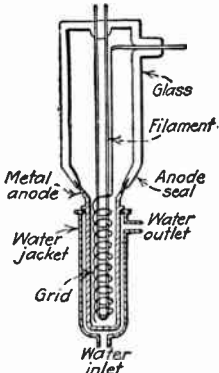


Fig. 260.—Schematic plan of high-power water-cooled vacuum tube for transmitting.

A water-cooling system is provided for properly cooling the four high-power amplifier tubes. This system is illustrated in Fig. 259, a water-cooled tube being shown in Fig. 260. The operation of this system is briefly as follows.

The water is circulated by a motor-driven centrifugal pump at the rate of approximately 50 gal. of water per minute. The moment the pump begins operation the water is forced through all the pipes, cooling radiators, water hoses, and tube jackets, and after completing its circulation through these parts returns to the drain tank. In view of the fact that the intense heat produced by the tube plates is slightly over 100 kw. of power, the hot water passes through a cooling radiator in which the heat is removed from the water and blown or radiated into free air. This radiator and its centrifugal blower are generally located out of doors and so designed that air is forced through the radiator pipes at the rate of approximately 15,000 cu. ft. per minute. The heat developed by the water after it passes out of each tube is recorded by dial-indicating thermometers in order that the proper temperature may at all times be ascertained.

The carrier frequency is generated by a crystal-controlled oscillator which delivers its output through six successive stages of radio-frequency amplification, the audio-frequency voltages being applied into the fourth stage through the medium of two low-frequency amplifying stages. The modulation of the radio-frequencies in the fourth stage (low-level modulation) is then amplified linearly through a class B push-pull amplifier and fed into the antenna by a transmission line.

Complete 100 per cent modulation is obtained over a frequency spectrum of from 30 to 10,000 cycles which produces high-fidelity transmission.

CHAPTER X

COMMUNICATION TRANSMITTERS

Radio transmitters designed for modern shipboard installation are primarily vacuum-tube¹ telegraph transmitters designed to emit continuous and interrupted continuous waves. Certain types of marine transmitters are combination telephone and telegraph transmitters, notable among this type being sets designed for the Coast Guard. A radio-telephone attachment has been designed for use with a conventional ship transmitter and will be described in this chapter.

I. MARINE TRANSMITTERS

250. Marine transmitters are very ruggedly designed to withstand hard usage and to reduce the possibility of breakdown. Their size is such as to permit passage through bulkhead doors and installation in small cabin rooms. Metal parts, as far as possible, are finished to resist attack by the salt air. Live circuits are protected against accidental contact by the operating personnel during rough weather by enclosing them as far as possible in protective screens and shields. Entry doors to the interior of the transmitters are sometimes fitted with interlocks which cut off high voltages when the door is opened for adjusting or other purposes.

It speeds up operation in marine work if the transmitters are fitted with break-in systems, and many transmitters are so fitted. Such a system permits the operator to hear signals when his key is up, just as anyone talking on a conventional telephone can receive while transmitting or talking.

Marine transmitters are designed to operate from the ship's power which is always direct current, usually 110 volts. In addition, certain low-power and emergency transmitters are designed to operate from a 12- or 32-volt storage battery in which case no motor starter is necessary, the motor being switched directly to the battery.

Frequency stability in a marine transmitter, while important, does not *necessitate* crystal control, and for this reason many transmitters

¹ The lack of space here does not permit descriptions of antiquated arc and spark transmitters and crystal detectors. Readers desiring information on these equipments are referred to the authors' "Radio Operating Questions and Answers," or to the authors' "Practical Radio Telegraphy," published in 1928, or to any of the older textbooks on radio.

designed for vessels of the merchant marine are master-oscillator power-amplifier transmitters, although some crystal-controlled transmitters are also used. Harmonics are suppressed by enclosing oscillating and radiating circuits in shields, and by interposing tuned circuits between the oscillator and the antenna. Parasitic oscillations are suppressed by utilizing a minimum number of tubes for the power required, and by the use of grid chokes on all tubes.

The law requires that certain vessels be equipped with emergency transmitters capable of transmitting over a distance of 100 miles, day or night. Such a transmitter is described in this chapter; it is the type ET-3650 and may well be studied first, as it utilizes a simple circuit without the refinements found on larger sets.

The power of a given transmitter is more or less fixed except for test purposes of short time duration when low power is used.

251. Keying Systems.—In order to form the wave emissions from a radio-telegraph transmitter to conform to the characters of the International Morse Telegraph Code it is necessary to start and stop these emissions at will. This is known as *keying* the transmitter and requires special circuit arrangements which must be designed as an integral part of the transmitter.

There are many systems of keying used to meet various conditions. The simplest system perhaps is that employed in the emergency transmitter type ET-3650, where the key is inserted in the primary lead of the plate transformer. This system is well adapted for this transmitter because it uses raw alternating current on the tube plates with its consequently modulated note, and also because only low power is handled. This method is unsuited to higher-powered transmitters using pure direct current on the plates of the tubes because of the e.m.f. inertia in the primary circuit of the transformer which would cause a chirping note to be emitted, and also because the key, in this position, must handle the total current drawn by the transmitter. This method, therefore, is not used except under the conditions encountered in a transmitter such as the type ET-3650.

The next method, in the order of simplicity, is perhaps that which causes the key to remove a blocking bias from the oscillator tube. A schematic diagram of this method is shown, Fig. 261(a). When the key is up, the tube is blocked due to the grid being negative to the filament by the amount of voltage drop across the resistance *AB*. When the key is down, the tube operates because the grid is at the same potential as the filament, minus only the normal operating bias caused by the drop across the grid leak.

An extension of the method just described is to apply the blocking bias to all the tubes, amplifiers as well as the oscillator. The general scheme is the same as that shown, Fig. 261(a), in which diagram the tube

shown can represent all the tubes in the transmitter. This method is employed in the ET-3626-C transmitter later described.

Another system of keying employed on the Mackay radio-transmitter type 101-A is the grid-choke method. In this system a choke heavy enough to stop oscillation of the tube is placed in the grid of the oscillator tube and is known as the *keying choke*. This choke is short-circuited when the key or keying-relay contacts are closed which permits the tube to oscillate; hence keying is accomplished.

In connection with crystal-controlled transmitters, it is generally undesirable to key the oscillator or the buffer amplifier, as to do so would disturb the frequency stability of the transmitter. The practice with these transmitters is to key the stages following the buffer amplifier which

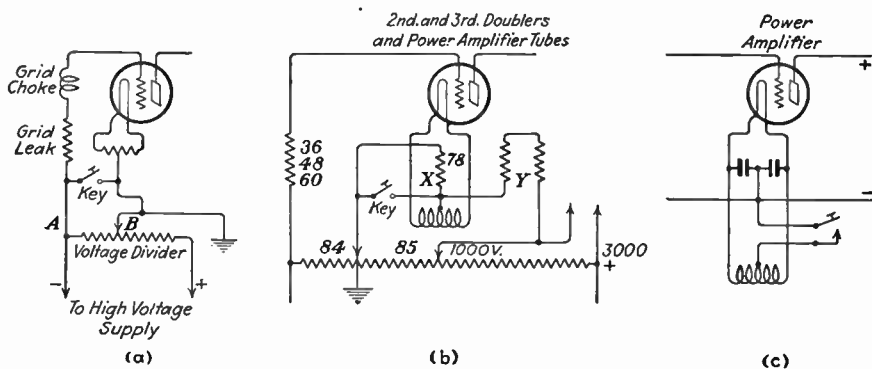


FIG. 261.—Various methods of keying a transmitter.

method is employed in the R.M.C.A. transmitter type ET-3656-A which is a high-frequency transmitter. This keying circuit is shown in the schematic diagram of Figs. 261(b) and 268, and its operation is further explained in Sec. 254.

252. Production of Interrupted Continuous Waves.—Interrupted continuous waves (*ICW*) are desirable in marine communication because of their greater attention-getting value when standby receivers are used. In other words, the receiver does not have to be tuned so close to the emitted wave in order for an audible signal to get through to the headphones when *ICW* is used. There are two general methods of interrupting the continuous wave generated by the oscillator, namely: (a) by means of a chopper; (b) by means of an audio oscillator. Strictly speaking only the chopper interrupts the wave, whereas the audio oscillator modulates the wave. For this reason, when the audio oscillator is used, the emitted wave form is sometimes referred to as *MCW*, meaning modulated continuous waves. The final effect, however, is the same, in that the emitted wave is broadened to cover a wider frequency spectrum.

a. In earlier days of radio detectors it was necessary to

when most ships were fitted with crystal interrupt the continuous-wave emissions in order that they could be made audible by a non-oscillating (crystal) detector. This was usually effected in arc transmitters, which were the only continuous-wave (CW) transmitters in common use, by a chopper in series with the antenna circuit. Likewise, any vacuum-tube transmitters operating with a chopper or

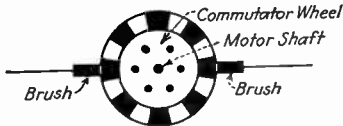


FIG. 262.—Chopper wheel and contact brushes. Black segments are conducting; white, non-conducting.

audio oscillator to produce ICW may be heard on a crystal- or other non-oscillating-detector receiver.

The earliest forms of vacuum-tube transmitters which were converted from spark sets required no device for interrupting the wave emission, as this emission was raw alternating current produced by self-excited transmitters using alternating current for plate power. Later models of tube transmitters, which were designed as such, required some method for breaking up the CW emission produced when direct current was used on the plates. A chopper designed in accordance with the scheme shown in Fig. 262 was used. This chopper was nothing more than a motor-

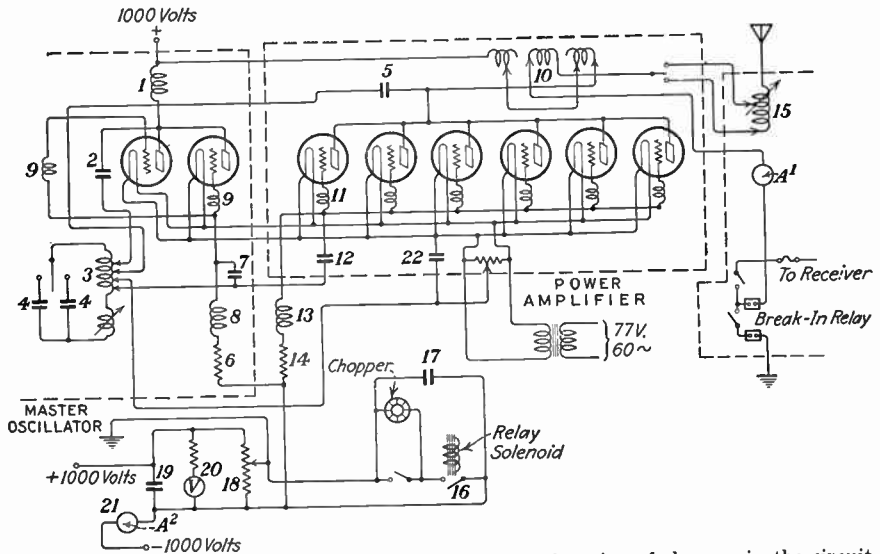


FIG. 263.—R.M.C.A. transmitter type 3626-A, showing location of chopper in the circuit.

driven commutator connected in series with the grid circuit of the oscillator and amplifiers. As the grid circuit is interrupted, so is the emitted wave. A chopper is used in the R.M.C.A. transmitter type 3626-A which is shown in Fig. 263. This transmitter has been superseded by later models utilizing an audio oscillator to produce the ICW.

b. Modern practice is to produce *ICW* by means of an audio oscillator as will be seen by studying such transmitters as the type ET-3626-C (Sec. 255) explained later in this chapter. A schematic diagram of an audio oscillator is given in Fig. 264. By studying the circuit, it will be noted that the tube is made to oscillate by feeding back the output from

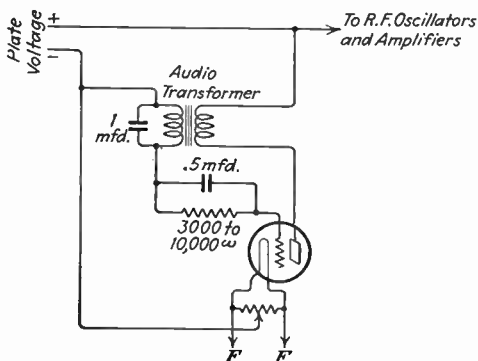


Fig. 264.—Audio-oscillator circuit for production of *ICW*.

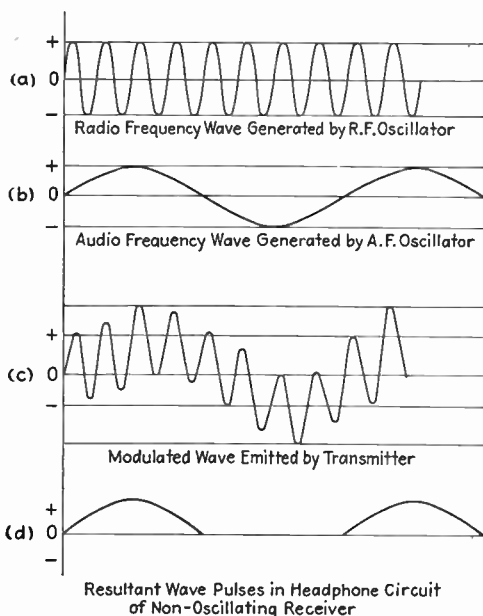


Fig. 265.—Curves showing action of audio oscillator on continuous wave emission.

the plate circuit to the input or grid circuit. In this respect the audio oscillator is not different from the radio-frequency oscillator. The audio oscillator, however, utilizes heavier condensers and larger inductance values. As the frequency of any vacuum-tube oscillator depends on the inductance and capacity in the plate and grid circuits, and, as the fre-

quency generated decreases as the value of these constants is increased, large values of inductance and capacity are used in the audio oscillator. In order to get the required inductance values it is necessary to use an iron-core transformer.

The theoretical action of the audio oscillator on the continuous wave is shown by the family of curves in Fig. 265. Curve (a) represents the continuous wave generated by the radio-frequency oscillator of the transmitter; curve (b) is the audio-frequency sine wave generated by the audio oscillator and coupled directly to the plate supply of the radio-frequency oscillator and amplifiers. The result is that the audio-frequency wave either adds to or subtracts from the value of the radio-frequency wave, resulting in a modulated wave as shown in curve (c). Curve (d) shows the resultant current in the headphone circuit of a receiver using a non-oscillating detector. Notice that the positive pulses of curves (b) and (d) are identical, and that for this reason the received note corresponds to the frequency of the audio oscillator used to produce the *ICW*. These waves are also known as *tone-modulated waves*. Antenna current, and consequently transmission range, is generally somewhat less when *ICW* is used than when *CW* is used.

253. Emergency Transmitter : R. M. C. A. TYPE ET-3650. *a. Description.*—This transmitter is designed primarily for installation on vessels where a low-power emergency transmitter is required. This set is designed to provide only *ACW* 700-cycle telegraphic operation on any frequency in the 600- to 800-meter band. It is designed to operate into an antenna having an effective resistance from 4 to 10 ohms, an effective capacitance of 600 to 1,400 mmf., and a natural wave length of from 225 to 450 meters. The transmitter is also suitable for use on antennas having an effective resistance from 2 to 12 ohms, but the output will be slightly lower when the antenna resistance is below 4 or above 10 ohms.

The transmitter will deliver 25 to 50 watts of alternating continuous waves (*ACW*, 700 cycles) radio-frequency power into the antenna just specified when operating on any frequency in the 600- to 800-meter band.

The transmitter uses four UX-210 radiotrons as power oscillators in a self-excited circuit, and the equipment has been designed to operate on a power supply of 12, 32, or 110 volts direct current. A photograph of this transmitter is shown in Fig. 266, a schematic diagram in Fig. 267.

b. Keying.—The set is keyed by means of an external key which closes the primary circuit of the plate transformer.

c. Adjusting Transmitter to Frequency.—The routine which follows applies in particular to the type ET-3650 transmitter and in general to all transmitters in which the oscillator is coupled to the antenna direct without a power amplifier.

1. Place the adjustable clip on inductance (5) to an approximate position on the coil. Place wave-change switch in proper position, if supplied.
2. Loosen antenna coupling as much as possible.

3. Put power-control switch in Low position and start motor generator.
4. Adjust filament voltage to proper value.
5. Close telegraph key.
6. Measure frequency of oscillator tank circuit by bringing wavemeter into inductive relation with this circuit, being careful not to bring it so close that damage to the meter results.
7. If the frequency is found to be too high or too low, the transmitter should be shut down while the adjustable clip on the inductance (5) is relocated on another turn or portion of a turn, either increasing or decreasing the inductance, depending on whether the frequency has to be increased or decreased.
8. Start transmitter and measure frequency again. Repeat this operation until correct frequency is obtained.

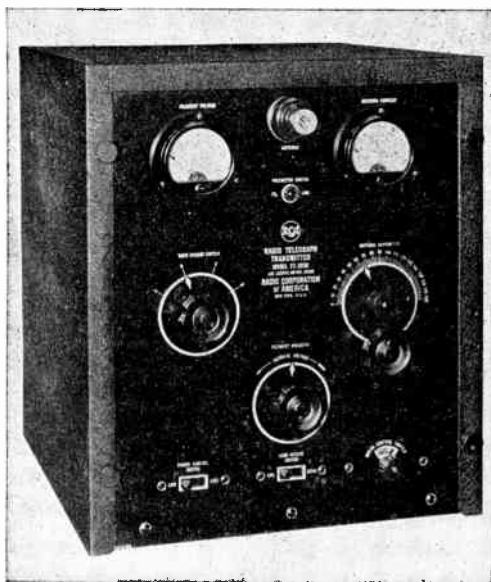


Fig. 266.—R.M.C.A. Type ET-3650 emergency transmitter. (Courtesy of R.M.C.A.)

9. Resonate the antenna circuit by selecting an approximate turn on the antenna variometer (4), and connect the lead going to the antenna to it. When the proper turn is located, a maximum reading will be obtained on the antenna ammeter as the antenna-inductance variometer is rotated through the resonant point of the frequency to which the tank or oscillator circuit is tuned. Continue adjusting turns until a point is found which will place the resonant point within range of the variometer.

10. Increase antenna coupling by bringing the antenna coil (3) closer to tank-coupling inductance (2) and lock the coil in position at a point just below where the antenna current drops sharply when the antenna variometer is rotated through resonance.

11. Place any shields or cover on the transmitter, and again check the frequency by coupling the wavemeter to the antenna lead. If the frequency is slightly high or low, increase or decrease the tank inductance (5) and recheck the frequency. The effect of shields and the position of the antenna-coupling coil will affect the emitted frequency slightly, and, therefore, this final adjustment is required.

12. A similar procedure is followed to calibrate to other frequencies which may be provided for through a wave-change switch.

13. Always tune the transmitter on low power and do not put on high power until circuits are in resonance.

Precaution: The operator should always be careful to remove plate power before any adjustments are made on the transmitter, unless these adjustments are made from switches or knobs on the panel.

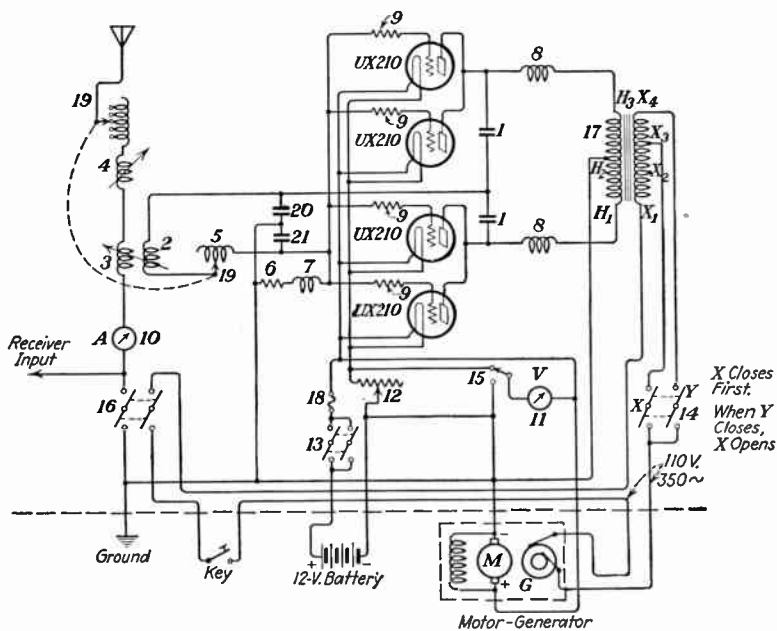


FIG. 267.—Emergency transmitter; R.M.C.A. Type ET-3650. (Courtesy of R.M.C.A.)

PARTS OF EMERGENCY TRANSMITTER (SEE FIG. 267)

- | | |
|--|---|
| 1. Plate-blocking capacitors. | 11. Filament voltmeter, direct current 0 to 15 volts. |
| 2. Tank-coupling coil. | 12. Filament rheostat, $1\frac{1}{2}$ ohms, 5 amp. |
| 3. Antenna-coupling coil. | 13. Motor-control switch. |
| 4. Antenna variometer. | 14. Power-control switch. |
| 5. Oscillator tank coil. | 15. Filament-voltmeter switch. |
| 6. Grid leak, 600 ohms, 60 watts. | 16. Send-receive switch. |
| 7. Grid choke. | 17. Plate transformer. |
| 8. Plate-choke coils. | 18. Line fuse, 60 amp. |
| 9. Grid parasitic resistors, 15 ohms each. | 19. Wave-change switch. |
| 10. Antenna ammeter, radio frequency 0 to 5 amp. | 20. Plate-tank capacitor. |
| | 21. Grid-tank capacitor. |

254. High-frequency Transmitter: R.M.C.A. TYPE ET-3656-A.

a. Description.—This transmitter is designed to provide continuous-wave telegraphic communication on any frequency from 6,000 to 20,000 kc./s. (50 to 15 meters) or from 5,550 to 17,150 kc./s. (54 to $17\frac{1}{2}$ meters), depending upon the model supplied. A schematic diagram of this transmitter is shown in Fig. 268.

b. Antenna Characteristics.—A single-wire antenna suspended vertically, and as free from miscellaneous guy wires and masts as possible, will be found to give very satisfactory results. The antenna should be a five-fourths-wave antenna, as measured from the antenna-connection post on the transmitter to the free end, and as referred to the operating harmonic it is desired to use. A five-fourths-wave antenna, resonant to 17,050 kc./s., will be approximately 58 ft. in length. If possible, another antenna of the vertical type but about 87 ft. long should be available, as it will make for better operating conditions on certain frequencies.

c. Notes on Operation.—A buffer-amplifier stage is used to follow the crystal oscillator. This stage serves to permit working the crystal-oscillator circuit lightly and to prevent reactions caused by tuning or keying the following amplifier stages. The buffer-tank circuit (18, 19) is tuned to the oscillator frequency.

Variable inductance (30), fixed inductance (30A) and fixed capacitor (29) form the tuned tank circuit of the first doubler stage, and this circuit is always tuned to twice the crystal frequency desired.

Variable inductance (44) and fixed capacitor (43) form the tuned tank circuit of the second doubler, and this circuit always tunes to four times the crystal frequency.

Variable inductance (56) and fixed capacitor (55) form the tuned tank circuit of the third doubler, and this stage tunes to four or eight times the crystal frequency depending on the output frequency desired.

Variable inductance (70), together with the tube and circuit distributed capacities, forms the tuned tank circuit of the power-amplifier stage, and this circuit is always tuned to the same frequency as the third doubler. In order to cover the entire frequency range effectively without necessitating a variable inductance (70) of too high a ratio, a capacitor (69) has been added. This capacitor is automatically connected in the circuit, for use when the lower frequencies are employed, by the band-change switch (69A) as shown.

The transmitter may be adapted to "current-feed" the antenna with the inductance (72) and capacitor (74) in series [switch (73) in position 2], or with the inductance (72) alone [switch (73) in position 3], or the "voltage-feed" system may be used by throwing switch (73) to position 1, thus eliminating the circuit to ground through the inductance (72).

Should it be desired to couple into a balanced transmission line for feeding an antenna or power amplifier the three-position switch (73) is placed in position 2, and the antenna selector switch (75) is thrown to the left (this is the "up" position on the panel). With the switches in these positions, voltage for driving the transmission line is applied through the antenna-coupling capacitor (71). The variable capacitor (74) and variable inductance (72) together form a tank circuit which is coupled

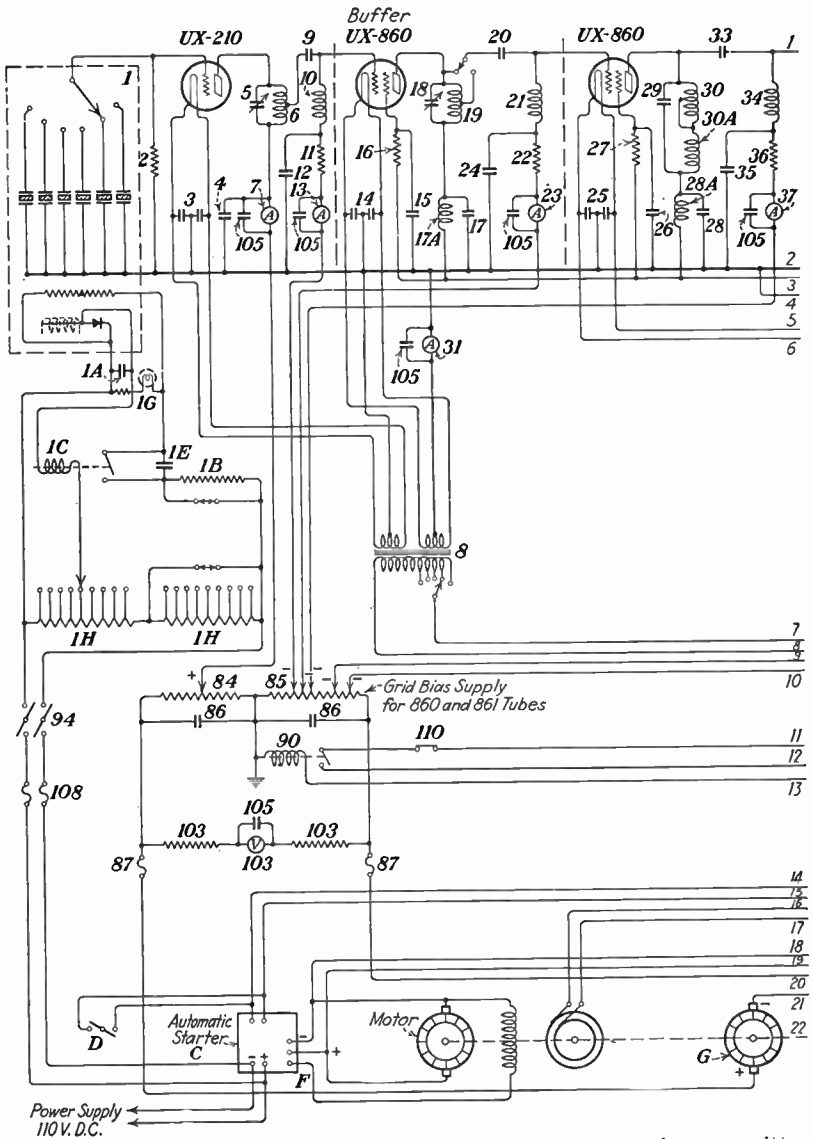
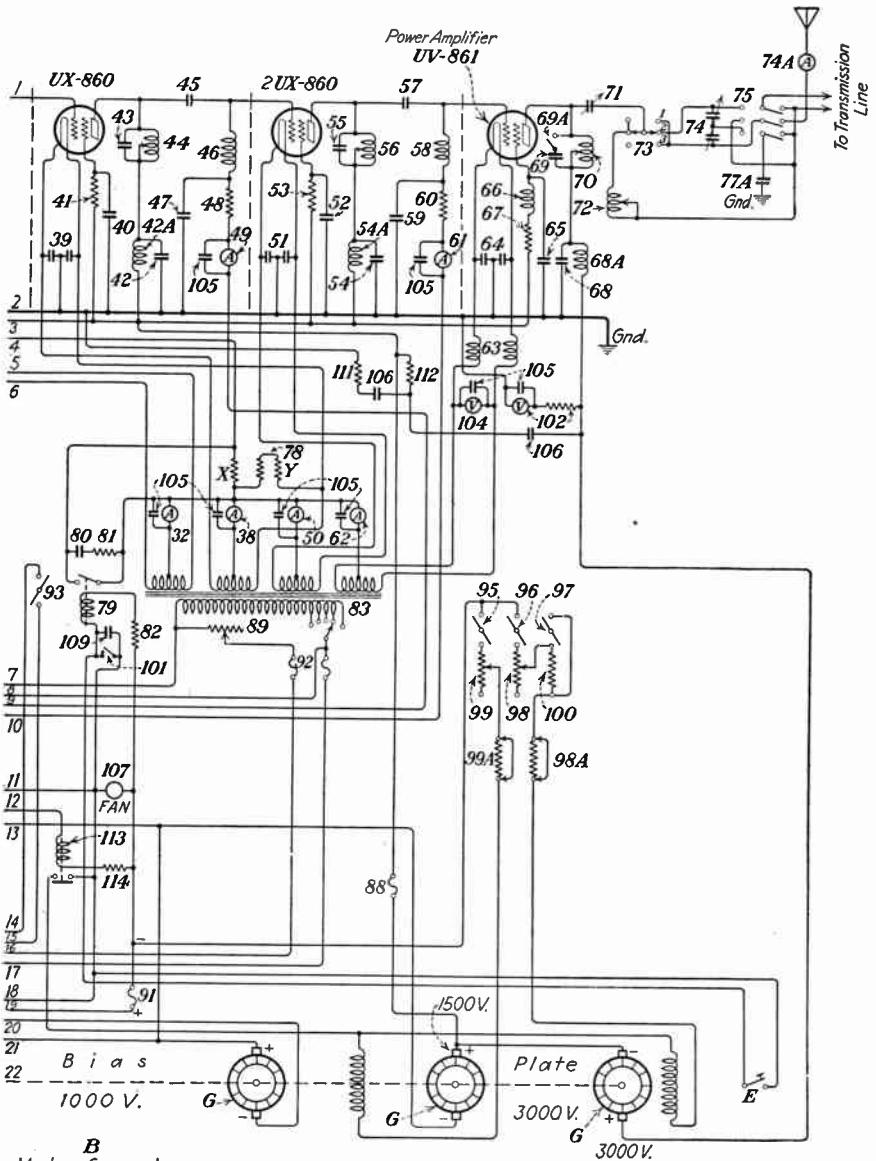


FIG. 268.—High-frequency marine transmitter;



B
 Motor-Generator
 R.M.C.A. type ET-3656-A. (Courtesy of R.M.C.A.)

directly to the transmission line. The transmission-line grounding through capacitor (77A) maintains the center of the variable capacitor (74) at ground potential.

d. Keying.—The keying of this transmitter is accomplished by keying all stages, except the crystal oscillator and the buffer, simultaneously. The crystal oscillator tube is allowed to run continuously for stability, and the stages following the first doubler are biased to cut-off so that when their excitation is removed, their plate current falls to zero. The keying scheme is shown in the schematic diagram Fig. 261(b) which is labeled to conform to the transmitter. The explanation of the keying which follows may be understood readily by referring to this smaller diagram and then comparing it with the large schematic of the entire transmitter. An inspection of the diagrams will show that keying is accomplished by inserting a high resistance (78x) in the path of the negative high-voltage return to the filament. Any plate current which flows with the key-relay contacts open must then flow through this high resistance. A small amount of current is maintained in the resistance by a part of the resistor section (78Y) which connects to the plate-voltage (+) supply for the tubes. When the key relay is opened, a portion of the voltage drop across the resistor (78X) is applied to the filament, which brings it to a certain positive value above the grid potential giving the grid a negative potential or bias with respect to the filament. This condition still further tends to reduce any plate current through the tubes to zero, and there is no radio-frequency output from the tubes.

In keying the transmitter, the keying relay short-circuits the high-resistance path between negative high-voltage (ground) and filament, thus removing the high bias from the grids and allowing the plate current to flow. Excitation is then supplied to the power amplifier, and radio-frequency output is obtained. Test key (101) and the telegraph key (E) on the operator's control table are wired in parallel, and they control the key relay (79). Capacitor (80) and resistance (81) serve to prevent any sparking at the keying-relay contacts. Resistor (82) is the key-relay series resistor.

e. Crystal-box Heater.—A thermostat located in the crystal box and which can be adjusted by turning a dial on the front of the panel, operates the heater relay (1C) thereby maintaining the heat in the crystal box to an approximate constant temperature of 50°C. A pilot lamp (1G) indicates when the crystal heater is on.

f. Tuning the Transmitter.—The following procedure is recommended for tuning the transmitter to any desired frequency within its range. Let us assume that the desired frequency is 18,100 kc./s. (16.57 meters).

1. Several hours prior to operation, close crystal-heater switch (94).
2. Start motor generator.
3. Adjust filament voltage to 11.0 volts.
4. Select the 2,262.5 kc./s. crystal by means of the crystal-selector switch. Crystal frequency is one-eighth of output frequency.
5. Apply bias voltage and adjust to 800 volts, not more.
6. Apply plate voltage and adjust to 3,000 volts. The tuning-control switch should be in OPERATE position until this adjustment is made, then throw to TUNE position.
7. Tune oscillator tank until crystal oscillates, which will be indicated by dip in oscillator plate current, and reading on buffer grid meter. Adjust to slightly below maximum grid reading.

PART OF R.M.C.A. TYPE ET-3656-A TRANSMITTER (SEE FIG. 268)

1. Crystal box.
- 1A. Thermostat capacitor.
- 1B. Heater series resistor, 600 ohms.
- 1C. Heater relay.
- 1E. Heater-relay capacitor, 0.5 mf., 600 v. d.c.
- 1G. Thermostat-indicator lamp.
- 1H. Heater-relay potentiometer, two 500 ω units.
2. Oscillator grid leak, 0.5 megohm.
3. Oscillator filament by-pass condenser, 0.1-0.1 mf., 250 volts, 15 amp. at 1,000 meters.
4. Osc. plate, by-pass cond., 0.005 mf., 2000 v. eff.
5. Oscillator tank condenser.
6. Oscillator tank inductance.
7. Oscillator plate ammeter, 0-100 ma.
8. Oscillator and buffer filament transformer.
9. Buffer coupling condenser, 0.002 mf.
10. Buffer grid choke.
11. Buffer grid leak, 3,000 ohms.
12. Buffer grid by-pass condenser, 0.008 mf.
13. Buffer grid ammeter, 0-10 ma.
14. Buffer filament, by-pass condenser, same as 3.
15. Buffer screen-grid, by-pass condenser, 0.1 mf., 2,000 volts, 20 amp. at 20,000 kc./s.
16. Buffer screen-grid resistor, 20,000 ohms (two).
17. Buffer plate by-pass condenser; same as 9.
- 17A. Buffer plate choke.
18. Buffer tank condenser.
19. Buffer tank inductance.
20. 1st doubler coupling condenser, 0.001 mf., 0.9 amp. at 1,000 meters.
21. 1st doubler grid choke.
22. 1st doubler grid leak.
23. 1st doubler grid ammeter; same as 7.
24. 1st doubler grid by-pass condenser; like 12.
25. 1st doubler filament by-pass condenser; like 3.
26. 1st doubler s.g. by-pass condenser; same as 15.
27. 1st doubler s.g. resistors, same as 16 (two).
28. 1st doubler plate by-pass condenser; like 15.
- 28A. 1st doubler plate choke.
29. 1st doubler tank condenser, 0.0002 mf., 5,000 volts, 6 amp. at 4,000 kc./s.
30. 1st doubler tank inductance.
- 30A. 1st doubler loading inductance.
31. Buffer plate ammeter, 0-100 ma., d.c.
32. 1st doubler plate ammeter, 0-300 ma., d.c.
33. 2d doubler coupling condenser; same as 20.
34. 2d doubler grid choke.
35. 2d doubler grid by-pass condenser; same as 12.
36. 2d doubler grid leak; same as 11.
37. 2d doubler grid ammeter; same as 7.
38. 2d doubler plate ammeter; same as 32.
39. 2d doubler filament by-pass cond.; like 3.
40. 2d doubler s.g., by-pass cond.; like 15.
41. 2d double screen-grid resistors; same as 16.
42. 2d doubler plate by-pass condenser; like 15.
- 42A. 2d double plate choke.
43. 2d doubler tank condenser, 0.00005 mf., 3,000 volts effective, 3.5 amp., at 4,000-10,000 kc./s.
44. 2d doubler tank inductance.
45. 3d doubler coupling condenser; same as 20.
46. 3d doubler grid choke.
47. 3d doubler grid, by-pass condenser; same as 12.
48. 3d doubler grid leak resistor; same as 11.
49. 3d doubler grid ammeter; same as 7.
50. 3d doubler plate ammeter, 0-1 amp.
51. 3d doubler filament by-pass cond.; like 3.
52. 3d doubler s.g. by-pass cond.; like 15.
53. 3d doubler screen-grid resistor, 20,000 ohms.
54. 3d doubler plate by-pass condenser; like 15.
- 54A. 3d doubler plate choke.
55. 3d doubler tank condenser, 0.00025 mf., 10,000 v. eff., 1 amp. at 1000 kc./s. (two).
56. 3d doubler tank inductance.
57. P.A. coupling condenser; same as 20.
58. P.A. grid choke.
59. P.A. grid by-pass condenser, 0.02 mf., 1,500 volts effective, 10 amp. at 1,000 kc./s.
60. P.A. grid leak, 3,000 ohms.
61. P.A. grid ammeter, 0-200 ma.
62. P.A. plate ammeter, 0-2 amp.
63. P.A. filament chokes (two used).
64. P.A. filament by-pass condenser; same as 3.
65. P.A. screen-grid by-pass condenser, 0.1 mf., 2,000 v. eff., 20 amp. at 4000-20,000 kc./s.
66. P.A. screen-grid choke.
67. P.A. screen-grid resistor; same as 53.
68. P.A. plate by-pass condenser, 0.02 mf., 4,000 volts effective, 15 amp. at 4000-20,000 kc./s.
- 68A. P.A. plate choke.
69. P.A. tank condenser, 0.00015 mf., 6,000 volts effective, 12 amp. at 4,000-5,000 kc./s.
- 69A. P.A. band-change switch.
70. P.A. tank inductance.
71. Antenna (Ant.) coupling capacitor.
72. Ant. tuning inductance.
73. Ant. tuning switch.
74. Ant. series condenser.
- 74A. Ant. r.-f. ammeter, (0-10) with thermo-couple.
75. Ant. selector switch.
- 77A. Transmission-line grounding cond., 0.002 mf., 6,000 v. eff., 7.5 amp. at 1,000 kc./s.
78. Keying resistor.
79. Key relay.
80. Key-relay condenser, 1 mf., 600 v., d.c.
81. Key-relay condenser series resistor, 10 ohms.
82. Key-relay series resistor, 400 ohms.
83. Main filament transformer.
84. Osc. plate potentiometer (voltage divider).
85. Bias potentiometer (voltage divider).
86. Oscillator plate and bias-filter condenser, 1 mf. each (four used in parallel) 1,200 v., d.c.
87. Osc. plate and bias fuses, 0.5 amp., 1,500 v.
88. Plate fuse for doublers, 1 amp., 2,000 volts.
89. Filament rheostat, 90 ω , 26-2 amp., 250 v. (three plates, each of the above rating, used).
90. Overload relay.
91. Exciter fuses, 10 amp., 250 volts (two used).
92. Filament fuses; same as 91 (two used).
93. Start-stop switch.
94. Heater switch.
95. Bias-generator field switch.
96. Plate-generator field switch.
97. Tuning switch.
98. Plate-generator field rheostat.
- 98A. Not used.
99. Bias-generator field rheostat.
- 99A. Not used.
100. Tuning resistors (two used).
101. Key test.
102. Plate voltmeter, 0-4 kv.
103. Bias voltmeter, 0-1½ kv.
104. Filament voltmeter, 0-15 v., d.c.
105. Meter by-pass condenser, 0.008 mf. (14 used).
106. Plate-filter cond., 9.12 mf., 1,650 v., d.c. (two).
107. Six-inch electric fan.
108. Heater fuses, 5 amp. 250 volts (two used)
109. Key condenser, 1 mf., 600 volts.
110. Door interlocks (six used).
111. Plate-filter series resistor, 500 ohms.
112. Plate-filter bleeder resistor, 20,000 ohms.
113. Interlock contactor.
114. Interlock contactor resistor, 500 ohms.

8. Adjust buffer tuning for maximum reading of grid current on first doubler stage.

9. Depress key and note plate current on first doubler stage.

10. Adjust first doubler tuning for sharp decrease in its plate current and maximum reading of grid current on second doubler.

11. Adjust second doubler tuning for maximum grid current on third doubler stage.

12. Adjust third doubler tuning for maximum grid current on power-amplifier stage.

13. Adjust power-amplifier tank circuit to resonance with third doubler as indicated by dip in power-amplifier plate current. I_p should not exceed 200 to 250 ma.

14. Set antenna-coupling capacitor (71) at near minimum, place the antenna-tuning switch (73) in position 1, and throw antenna-selector switch (75) to the right (this is the DOWN position on the panel). Then adjust antenna-tuning inductor (72) for resonance which will be indicated by a reading of the antenna ammeter and an increase in power-amplifier current. Compensate for effect of antenna on power-amplifier plate current by readjusting power amplifier so that a minimum plate-current reading is maintained. This minimum will be higher when the antenna load is coupled to the tubes.

15. As soon as the antenna has been resonated, the tuning-control switch on the transmitter panel should be thrown to OPERATE position. This shorts out a resistance in the plate-generator field circuit and brings the plate voltage up to normal operating value. The setting of the antenna-coupling capacitor should next be increased until the power-amplifier plate current is approximately 850 ma., care being taken that the power-amplifier tank circuit is maintained in resonance.

Curves showing the approximate settings for the various controls are supplied by the manufacturer when the set is installed, and it should be a relatively simple operation to tune the set by following the approximate values given, making any finer adjustments necessary on the spot.

g. Adaptability of the Transmitter.—While this transmitter has been designed primarily for installation aboard ship, the features which make it preferable for this use are largely mechanical and not electrical. In other words, this transmitter can be supplied for installation ashore to operate from three-phase 220-volt 60-cycle, or 220-volt direct-current.

In so far as the theory of operation is concerned all high-frequency transmitters are very much alike in circuit detail and in tuning operations. Therefore, high-powered high-frequency transmitters, for example, like those used by RCA Communications, Inc., for transcontinental and transoceanic work, and rated at from 20 to 40 kw., are basically the same as this ship transmitter with the addition of the necessary power-amplifier stage rectifiers and other power-supplying equipment. The operations necessary to tune and operate such a high-power transmitter would be similar to the operations described for this ship transmitter, with the added operation of tuning a high-powered final power amplifier. The present power amplifier as shown in Fig. 268 then becomes the *exciter power amplifier* and is followed by the high-powered stage.

255. Medium-frequency Transmitter: R.M.C.A. TYPE ET-3626-C.

a. Description.—This transmitter is designed primarily for ship installa-

tion, but it may be operated on alternating current and used at shore stations if desired. This transmitter is especially interesting to the radio-telegraph operator as it is designed to be used with a radio-telephone attachment if desired. This telephone attachment, known as the model AT-829, will be described later. Many American vessels are fitted with this type transmitter, a wiring diagram of which is shown in Fig. 269.

b. General Explanation.—The outfit consists of three units, namely, the radio panel, the motor-generator set, and the operator's control unit. Two wave-length ranges are provided on this set as follows:

Range	Wave length, meters	Frequency, kilocycles	Power, watts	Emission
Lower.....	600- 800	500-375	500-750	<i>CW</i> and <i>ICW</i>
Upper.....	1,800-2,400	167-125	500	<i>CW</i>

This set uses a master-oscillator power-amplifier circuit, using the 50-watt, model UV-211 radiotron throughout. The oscillator consists of one tube connected in a Hartley circuit, and the amplifier consists of six tubes in parallel. Another tube is connected as an audio oscillator and modulates the *CW* emission when *ICW* operation is desired.

The filaments of the tubes are heated by alternating current supplied by the motor generator and taken off from one end by slip rings. This alternating current is supplied at 77 volts at 60 cycles. It is reduced to proper filament voltage by a rheostat and step-down transformer.

The motor generator supplies a plate potential of 1,000 volts. The driving motor can be furnished to operate on 32, 110, or 220 volts direct current, 110 or 220 volts, 60-cycle, single-phase, and 110, 220, or 440 volts, 60-cycle two- or three-phase alternating current.

Although not shown in the diagram, two entirely separate oscillator and antenna-tuning circuits are provided, one for the shorter intermediate wave lengths and one for the longer waves. A transfer switch is provided for changing from one to the other.

A switch is provided to select either the *CW* or *ICW* method of signaling. The audio oscillator is intended for use on the intermediate-wave range but may be used on the long waves if desired. Radiation is somewhat decreased when *ICW* is used. A three-point dial switch is connected in the audio-oscillator circuit and permits the selection of three different tone frequencies, 500, 700, or 1,000 cycles approximately. The modulated continuous wave emitted when the audio oscillator is in operation can be received by either an oscillating detector, such as the regenerative or superheterodyne detector, or by a non-oscillating detector such as a silicon or galena-crystal detector.¹ *ICW* emission is broader in

¹ The crystal detector is no longer used on American vessels but may still be in use to some extent on foreign ships, especially freighters (see note 1 at beginning of this chapter).

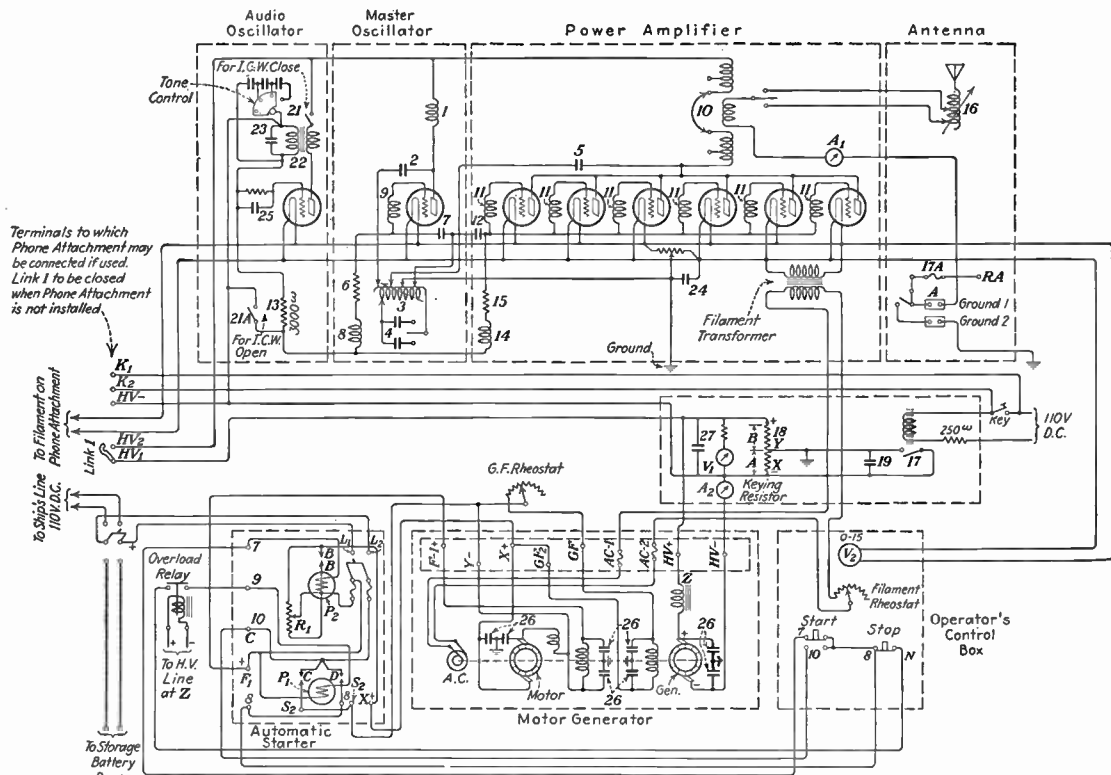


Fig. 269.—Medium-frequency marine transmitter; R.M.C.A. type ET-3626-C. (Courtesy of R.M.C.A.)

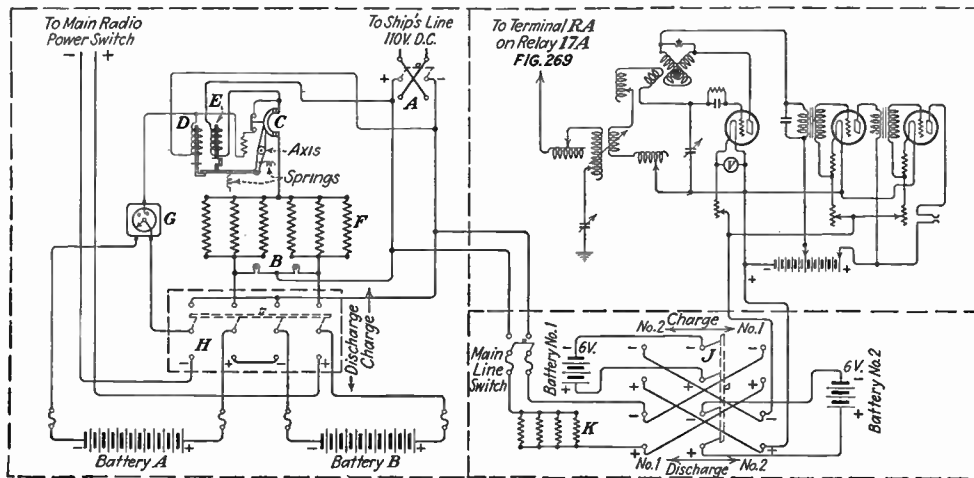


FIG. 270.—Auxiliary power supply and receiver usually used in conjunction with the R.M.C.A. transmitter ET-3626-C (Fig. 269). Figure 269 and this diagram, taken together, form a complete modern marine radio installation such as is used aboard American vessels.

A Polarity-reversing switch.
B Trickle-charge lamps.
C Circuit breaker.

EXIDE BATTERY SWITCHBOARD

D No-voltage release coil.
E Overload release coil.
F Main charging resistors.

G Ampere-hour meter.
H Charge-discharge change-over switch.

R.M.C.A. RECEIVER-BATTERY CHARGER

J Charge-discharge cycle switch. K Charging resistors.

R.M.C.A. RECEIVER (IP-501-A) (See Fig. 289 for details.)

PARTS OF R.M.C.A. TYPE ET-3626-C TRANSMITTER (See Fig. 269.)

1, 8, 14. 400-turn duolateral coil.

2, 7, 12. 0.004 mf. each.

4. { 0.002 mf., 5,000 volts for intermediate band.

5. { 0.004 mf., 5,000 volts for long-wave band.

6. 0.00014 mf., 6,000 volts.

13. 3,000-ohm grid leak.

15. 100-ohm grid leak.

18. Potentiometer, two 20,000-ohm resistors.

19. 1 mf., 750 volts.

23. Two 0.5 mf., 1,750 volts; one 1 mf., 1,750 volts.

24. 0.5 mf., 750 volts.

25. 3,000-ohm resistor, 0.5 mf., 1,750-volt condenser.

26. Protective condensers.

27. Main-filter condenser, 6 mf., 1,750-v.

A1. Antenna ammeter, 0-15 amp.

A2. Plate ammeter, 0-5 amp.

V1. Plate voltmeter, 0-15 volts.

V2. Filament voltmeter, 0-15 volts.

Note: Parts not listed here are explained under theory of operation of this transmitter, Sec. 255b, c, and d.

character than pure *CW* and is, therefore, used to a great extent for short-distance ship-to-shore and ship-to-ship communication as it permits the use of standby circuits in receivers and is generally more adaptable to this kind of work.

Although break-in operation is provided for through the keying relay, a separate receiving antenna may be used if desired.

For ship installations the motor generator is designed to operate from a direct-current supply. The motor is a four-pole machine for operation on 100 to 125 volts direct current. The motor is provided with slip rings and acts also as a rotary converter supplying current to the filaments at 77 volts, 60 cycles. The generator is a two-pole shunt-wound machine giving a direct-current potential of 1,000 to 1,200 volts at a speed of 1,750 r.p.m.

The operation of the set is controlled by the operator's control unit. This is a small metal box just large enough to contain a filament rheostat for controlling the filament voltage to all the tubes, a filament voltmeter, and the motor-generator *START* and *STOP* buttons. Additional control buttons may be installed at remote points if desired. This unit is placed on or near the operating desk.

c. Theory of Operation.—The functions of each piece of apparatus will now be traced referring to them by the numbers shown on the schematic diagram, Fig. 269.

Assuming that the direct-current line switches are in a closed position, the operation of the automatic starter circuit is as follows: When the *START* button is momentarily depressed the metal bar closes the contacts 10, 8, and *N* which connects one side of the plunger coil P_1 to the negative side of the line. Since the other side is permanently connected to the positive side of the line, the plunger solenoid becomes energized, causing the contacts *CD* to close. This excites both the motor and generator field windings simultaneously and also connected one side of the armature through the motor series field to the negative side of the line. At the instant that the *START* button was released, contact 7 connected with 8 which completes the circuit through the closed contact *D* and excites the plunger winding P_2 , causing it to rise slowly. A small current is flowing through the armature since the resistance R_1 is in series with it; so the motor starts slowly building up a counter electromotive force gradually. After several seconds the plunger P_2 causes the lower moving contact arm *B* to connect with the upper contact *B*, thereby short-circuiting the starting resistance R_1 and bringing the motor up to full speed.

When the *STOP* button is momentarily depressed, the contact *N* is broken and the circuit through the coil P_1 is opened. This breaks the contacts *C* and *D* which simultaneously disconnects the negative side of the supply line to the motor armature, motor, and generator fields. Plunger P_2 will also be released since the negative side of the line is now completely isolated from all equipment. The circuits will again be reestablished when the *START* button is depressed to make contact again with the negative side of the line at contact 10.

In the event that an overload or a short circuit should develop in the high-voltage circuit, the excessive current flow through the relay winding *Z* will open the circuit through P_1 and break the contacts *CD*, disengaging all motor and generator circuits.

The high-voltage current is fed to the plate of the oscillator tube through the radio-frequency choke coil (1), and the blocking condenser (2) separates the direct

current from the oscillating circuit. The variable inductance (3) and the condenser (4) constitute the oscillating circuit. The switch providing for choice of capacity (4) to conform to desired operating wavelength must be closed.

The proper adjustment of the plate, grid, and filament taps are made by the manufacturer for optimum efficiency and frequency stability and need not be adjusted after leaving the factory. The internal capacity of the amplifier tubes is prevented from affecting the oscillator by the neutralizing condenser (5). This condenser also performs the conventional function of preventing the amplifier from going into oscillation at any time.

The resistor (6) maintains the correct grid bias on the oscillator tube whereas the condenser (7) separates the d.c. grid current from the oscillator circuit. The choke coil (8) keeps the radio-frequency current on the grid from the keying circuit. The choke coil (9) is to prevent oscillations of a very short wave length (parasitics) from starting.

The high voltage is fed to the plates of the six amplifier tubes through the air-core transformer (10), thus eliminating the need for a plate choke and blocking condenser such as were used for the oscillator. The amplifier grids each have a choke coil (11) which has the same function as the choke coil (9) in the oscillator circuit.

The amplifier grids receive their excitation from an adjustable contact on the inductance (3). The condenser (12) blocks the direct-current grid-bias voltage, and is also the radio-frequency feed condenser to the P.A. grids. During *ICW* operation the resistor (13) prevents overloading the audio oscillator and acts as an additional grid-leak resistor for the power amplifier.

The inductance (14) keeps radio-frequency current on the grids of the amplifier tubes away from the keying circuit. The resistor (15) serves to keep a correct grid bias on the amplifier when the transmitter is in operation.

The variometer (16) controls the tuning of the antenna. The power in the antenna with a given plate voltage depends on the ratio between the primary and secondary turns in the transformer (10). This power may be increased by increasing the difference between the number of turns in the primary and secondary of this transformer.

The keying relay (17) is controlled by the hand key and obtains its current from the direct-current supply. A Leach relay, model 18, type S3 is used. This relay has auxiliary contacts (17A), shown in ground lead of antenna circuit, which close preceding the closing of the grid-circuit contacts and open after the opening of the grid-circuit contacts, thus providing break-in operation.

d. Keying is accomplished by biasing all tubes to the cut-off point when the key is up. This can most easily be understood by referring to the schematic diagram of the keying scheme, Fig. 261 (a). In the following analysis reference is made to Fig. 269.

When the keying contacts (17) are open, the plate current is forced to pass through the section of the resistor (18) extending from the negative terminal of this resistor to the tapped connection which is grounded. It is to be remembered that the electronic current in a metallic conductor flows from negative to positive (older theories to the contrary notwithstanding), and, therefore, the end of section *A*, which is connected to the negative line, is at a higher negative potential than is the center, *Y*, of this section. The grounded contact is adjusted so that there is about a 250-volt drop across section *A*, which means that the negative end of *A* marked *X* is 250 volts more negative than the other end marked *Y*. With the key up, the grids are connected to point *X* and they are, therefore, 250 volts negative in respect to point *Y* which is grounded. The filaments are center-tapped and grounded, which puts them at the potential of point *Y*. This makes the grids 250 volts negative in respect to the filament, and they are said to have a 250-volt negative bias, enough to block the tubes. When the keying contacts are closed, the grids are connected directly

to the ground (filaments) through their respective grid leaks, and normal operation takes place.

With this system of keying the high-voltage generator must supply the blocking-bias voltage in addition to the plate voltage. Therefore, if the plate voltage during operation is 1,000 volts and the blocking-bias voltage is 250 volts, the drop across resistor (18), or the output voltage of the generator, must be 1,250 volts. In one respect this extra voltage (250 volts) is lost as it cannot be used as plate voltage, but the advantages of this system in effective keying make it a desirable arrangement.

Arcing at the contacts of the keying relay (17) is minimized by the condenser (19). Condenser (27) acts as a radio-frequency by-pass around the high-voltage generator, prevents alternating-current surges from damaging the generator, and acts to reduce commutator ripple. Its capacity is 6 mf.

The audio oscillator used for *ICW* operation consists of one tube as shown in the diagram. During *ICW* operation, switch (21) is closed and switch (21A) is open. The tube receives its plate energy through the primary winding of the iron-core transformer (22). The secondary of this transformer is connected in series with the grid of the tube and couples the grid to the plate circuit of the tube thus causing it to oscillate. The condenser (23) across the secondary of the transformer provides a by-pass for the grid current around the winding and increases smoothness of operation of the audio oscillator. Condenser (24) is a radio-frequency by-pass around the filament center-tapped resistor. A negative operating bias is maintained on the audio-oscillator tube by the grid leak and condenser (25).

e. Adjusting the Transmitter.—In working with this set it must be constantly borne in mind that when the generator is running there is a high voltage on the plate circuit. *Contact with this circuit may cause death.* Therefore, **DO NOT TOUCH THE SET OR TUBES WHEN THE GENERATOR IS RUNNING. SHUT DOWN MOTOR GENERATOR WHEN MAKING ADJUSTMENTS.**

The initial adjustment of the set to the various frequencies on which it will operate is made by the installation engineer, and the various number of turns on the intermediate- and long-wave coils are noted on a card and posted near the transmitter for the information of any one concerned. Thereafter it should only be necessary to tune the set to maximum radiation.

f. Practical Operation.—To tune to maximum radiation proceed as follows:

1. Turn generator and filament rheostats to minimum.
2. Start motor generator.
3. Adjust filament voltage to 10 volts.
4. Adjust plate voltage to 1,000 volts.

5. Throw transfer switch to desired wave band, the exciter or oscillator already being tuned to the desired frequency. Throw switch to *CW*. Adjust proper antenna-tuning variometer until maximum radiation is indicated on the radiation meter, while pressing down the **TEST** button. When maximum radiation has been attained, the set is ready for operation. The telegraph key may now be manipulated for the transmission of messages.

6. To shut down set, press **STOP** button on operator's control panel.

7. To operate *ICW*, throw the "*CW-ICW*" switch to *ICW*, and select desired audio frequency by setting the three-point dial switch provided for this purpose to the proper position.

Caution: Do not touch tone-frequency switch when key is pressed as severe burns may result.

This set may be operated with as few as three amplifier tubes, if necessary, if full number is not available. If such operation is carried on, the plate voltage must be decreased to avoid overheating of the plates.

g. Antenna.—This transmitter is designed to operate with the conventional quarter-wave radiator generally installed aboard ship. The length of wire measured from the antenna terminal on the transmitter to the remote end of the antenna (L type) may be approximately from 250 to 350 ft., as explained in Sec. 181, Ship Antennas.

Generally an antenna having a capacity of 0.001 mf. and a radiation resistance of 6 ohms is used. When used with an antenna having these characteristics, the radio-frequency radiation will be on the 600- to 800-meter band from 9.8 to 10.2 amp., and on the 1,800- to 2,400-meter band from 9.6 to 10 amp. when the plate voltage is 1,000 volts.

256. Radiophone Attachment: R.M.C.A. MODEL AT-829.—This attachment is for use with the R.M.C.A. model ET-3626-A, -B, or -C radio-telegraph transmitters and makes possible voice as well as code transmissions. The radio-telegraph operator's knowledge of radio-telephone apparatus need not be so extensive as that of the broadcast operator, but it should be broad enough to include an understanding of radio telephony as used in the mobile service. A study of this attachment will give him this knowledge provided he understands the principles of radio telephony, modulation, and microphone operation particularly Chap. IV on Transmitting Circuit Principles and Secs. 25, 116, and 194.

a. General Description.—This attachment is mounted in a metal cabinet with a panel front and perforated metal sides. Its dimensions are $8\frac{1}{4}$ in. wide by $24\frac{3}{8}$ in. high by $30\frac{3}{16}$ in. deep. A small key switch for table mounting shifts from the operator's control microphone 2 to the remote microphone 1, and also provides for intercommunication between the microphones. One of the microphones is located in the radio room and the other may be located at any desirable remote location such as the pilot house. A circuit diagram of the complete attachment is shown in Fig. 271. This diagram is drawn so that it can be added to the diagram of the type ET-3626-C transmitter (Fig. 269) by connecting like-numbered terminals on both equipments together.

The model AT-829 telephone attachment modulates the output of the transmitter by the use of the constant-current system, using one tube (type UV-211) as a speech amplifier and four tubes of the same type as modulators.

A switch mounted on the panel and shown as a five-pole switch in the diagram, changes the necessary connections from telegraph to telephone. The *CW (Teleg.)* position completely disconnects the telephone attachment and connects HV_1 to HV_2 thus returning the high voltage directly to the transmitter.

To telephone, throw the panel switch to *PHONE* and press either microphone button. The break-in system is maintained in the operation of the radiophone, and it is only necessary to release the microphone button and listen for a reply, providing the receiver is properly tuned.

b. The Microphone System.—Current for the microphones is supplied by a 6-volt storage battery which also energizes a relay which closes the key circuit, thus operating the key relay on the main transmitter.

Either microphone may be thrown in by the operator's key switch. If the break-in system is to be used at both microphones, it is necessary to have a connection made to the radio receiver from the terminals marked *RXC* on the telephone-attachment panel. Separate headphones must be used for telegraph service as the ones furnished with the microphones must be left on the hook when not in use.

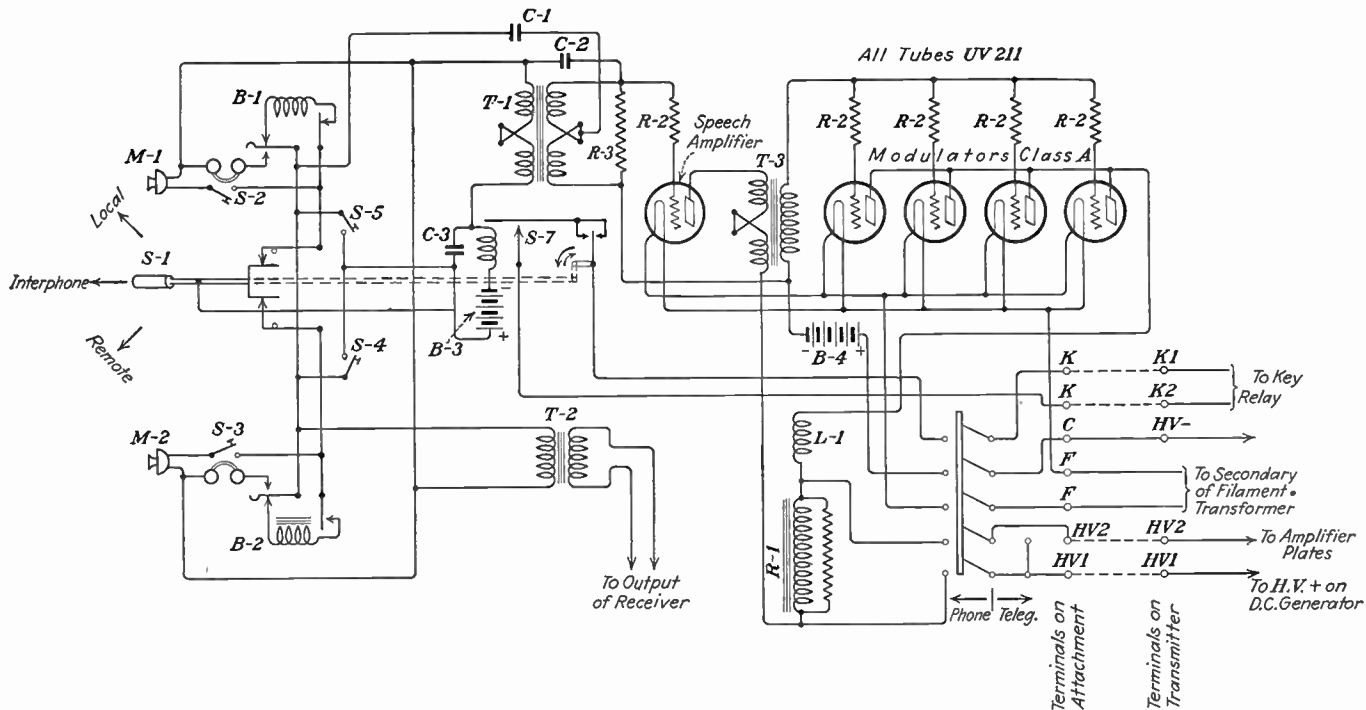


FIG. 271.—Radiotelephone attachment, R.M.C.A. Type AT-829, for use with R.M.C.A. type ET-3626-C transmitter.

PARTS OF RADIO-TELEPHONE ATTACHMENT R.M.C.A. TYPE AT-829

B_1 }	Buzzers.	M_2	Microphone 2, local.	S_4 }	Buzzer buttons.
B_2 }		R_1	Modulator-plate reactor.	S_5 }	
B_3	Microphone and relay battery.	R_2	Grid-leak resistor.	S_6	Key relay.
B_4	45-volt C battery.	R_3	Grid impedance.	S_7	Relay.
$C_{1,2,3}$	Condensers, 5 mf. each.	S_1	Key switch.	T_1	Microphone transformer.
L_1	Modulator-plate choke coil.	S_2 }	Push-to-talk buttons.	T_2	Receiver coupling transformer.
M_1	Microphone 1, remote.			T_3	Amplifier transformer.

The microphone-stem button must always be pressed down when speaking, whether in radio transmission or in intercommunication. A small pearl push-button is provided in the base of each microphone for ringing. These buttons are marked S_4 and S_5 on the diagram.

c. Installation.—The telephone attachment model AT-829 should be located, if possible, on the right side of the transmitter. If necessary, it may be located elsewhere in the radio room.

d. Adjustment.—After the set has been installed and connected, and the necessary tubes and batteries installed, the set is ready for adjusting. The B batteries furnish a 45-volt grid bias to the amplifier and modulator tubes which should make the plate current of the amplifier and modulator tubes about 75 ma.

It should be noted that the filaments of the telephone attachment are in parallel to the secondary of the filament transformer of the main transmitter which is shunted by a resistance which is center-tapped and grounded.

Throw the switch on the telephone panel to *CW* and see that the telegraph set operates normally. Then throw the telephone-panel switch to *PHONE* and adjust the filament voltage to 10 volts. Throw the key switch to microphone *M-2* and test it. The set should immediately draw plate current and put power into the antenna as shown by the radiation ammeter. When the microphone is spoken into, the plate current should increase slightly. If everything operates satisfactorily, the remote microphone *M-1* should be tested for intercommunication and modulation of the transmitter.

e. Operation.—In order to use the set as a radio telephone, after the preliminary adjustments have been made, it is only necessary to throw the switch on the attachment to *PHONE*. When operating phone, the filament voltmeter must read 10 volts.

To use the operator's microphone throw the key switch S_1 to *LOCAL*. Then push the microphone button and speak into the microphone. It is only necessary to release the microphone button to receive.

In order to talk to someone at the remote microphone from the operator's microphone, it is only necessary to throw the key switch to the center position which is marked *INTER-PHONE* and press the call button on the microphone stand.

If a party at the station with which communication is established wishes to talk to someone at the remote microphone proceed as directed in the preceding paragraph, and, when the party at the remote microphone answers, tell them to go ahead, and throw the key switch to *REMOTE*. The operator may then monitor the conversation to see that there is no difficulty due to inoperation of either set.

f. Maintenance.—Keep the set clean. Renew B batteries about every 6 months. Renew bad tubes when necessary.

If the set does not modulate, this may be due to a defective amplifier tube; try replacing it.

If the tubes get hot when the microphone button is not pressed, it is quite evident that the tubes are not receiving the cut-off bias from the main transmitter owing possibly to the inoperation of the relay, or to a tube which is shorted from grid to filament. The lead to C+ should be inspected to see that it is connected to the negative high-voltage terminal of the transmitter (*HV*—).

If the tubes heat too much when the button is down, it is a sign that the tubes are not blocking because they do not receive operating bias. The obvious remedy is to examine the C battery and connections and to test the individual batteries.

If necessary, the attachment may be operated with less than four modulator tubes, but the speech-amplifier tube must always be used.

257. Radio Telephone-telegraph Transmitter. *a. Description.*—The U.S. Coast Guard model T6A transmitter, shown in Fig. 272, is designed

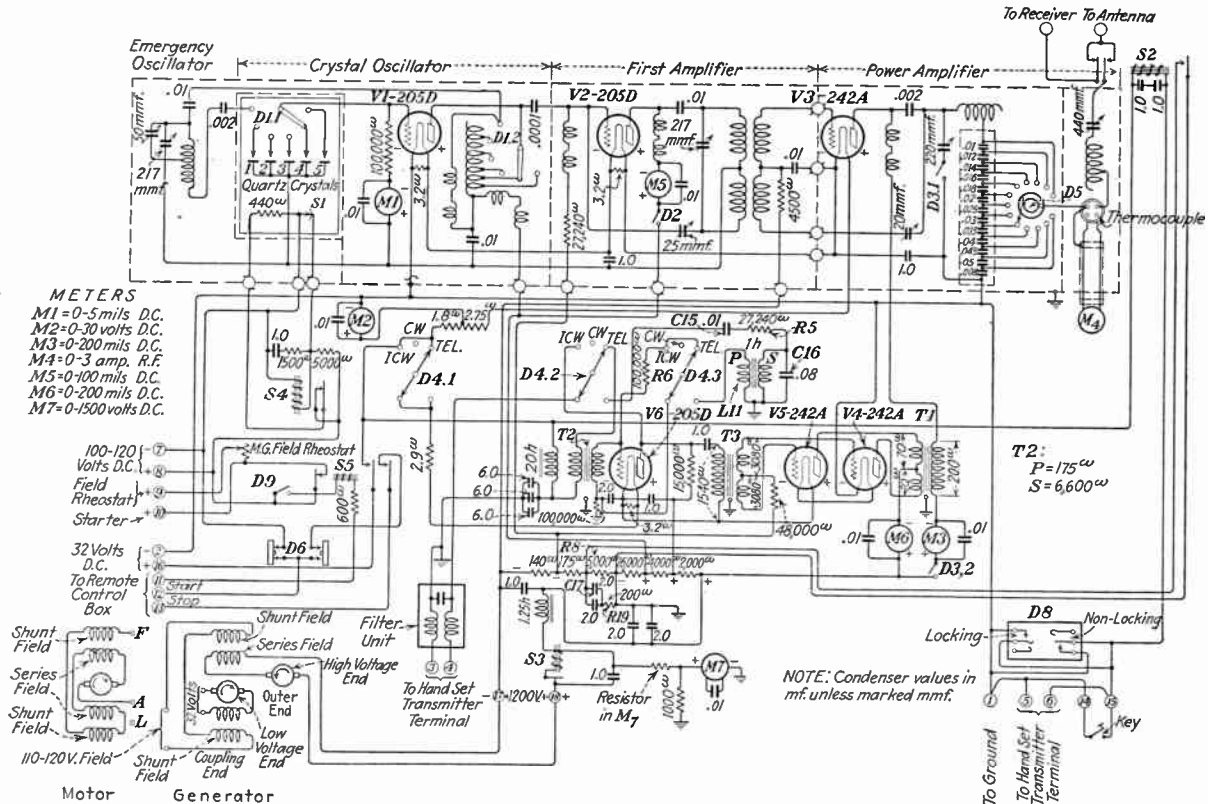


FIG. 272.—Radio telephone-telegraph transmitter. U. S. Coast Guard model T6A. (Courtesy of U. S. Coast Guard.)

to operate within the frequency band of 2,307 and 3,333 kc./s. (130 to 90 meters) and is capable of delivering not less than 35 watts of radio-frequency energy into an antenna having an effective capacity of 175 mmf. and a natural frequency of 3,120 kc./s. The effective resistance of this antenna should range from 14 ohms at 2,307 kc./s. to 31 ohms at 3,333 kc./s.

Most of the significant features of this transmitter become evident from a study of the schematic diagram. The transmitter is capable of complete modulation, thus making it possible to impart a maximum of signal energy to the carrier. A three-position switch provides for quick change to either CW, ICW, or PHONE. The oscillator crystals are arranged in a constant-temperature chamber. An emergency oscillator is provided for use in case of crystal failure or for operation on frequencies within the range of the transmitter, for which crystals are not provided.

This transmitter utilizes a modified Heising system of modulation. Link switch $D_{3.1}$ is provided to permit the insertion of a sensitive thermomilliammeter for use while neutralizing the transmitter.

For modulated continuous wave (ICW) transmission, the speech amplifier is converted into a 1,000-cycle oscillator. When in the ICW position, the switch $D_{4.3}$ connects a tuned circuit C_{16} - L_{11} to the grid circuit of V_6 . The switch $D_{4.2}$ connects the feed-back resistance R_6 and the capacity C_{15} from the plate to the grid circuit of V_6 permitting the circuit to oscillate. This audio-frequency oscillator then drives the modulator tubes V_4 and V_5 . For this type of transmission the resistor R_6 is retained in the grid circuit in series with the winding P of the coupling device L_{11} , since, by such construction, changes in the grid circuit of V_6 , due to switching, are minimized and R_6 can have no detrimental effects as its impedance is small compared to that of the winding P of the inductance L_{11} .

In the modulator circuit, the switch $D_{4.1}$ is arranged to disconnect the filaments of the speech amplifier to minimize current drain when using CW transmission. To assure high quality of transmission, the transmitter-carrier output power should not exceed 35 watts for telephon-operation.

Keying for either CW or ICW transmission is accomplished by short-circuiting the resistor R_8 with the contacts of relay S_2 . When R_8 is not short-circuited, a high negative bias is placed on the radio-frequency amplifier tubes preventing them from delivering power to the antenna.

The telegraph key controls the operation of the keying relay by closing the 32-volt generator circuit through the relay winding. For telephone transmission the lever on the hand-set transmitter controls the operation of the keying relay in the same manner. The spark-absorption circuit, C_{17} - R_{19} , is placed across the keying contacts of the relay to protect them against arcing.

b. *Tuning Adjustments.*—While the space is not available here to explain the tuning of this transmitter in detail, the advanced student or operator will understand this operation from the following tabulation of necessary operations.

1. Set the CRYSTAL-SELECTOR switch for the desired frequency.
2. Set each dial at the approximate tuning adjustment for the desired frequency as shown by tuning chart.
3. Set the ANTENNA-COUPLING switch on the proper point.
4. Place the CHANGE-OVER SWITCH in the TEL position.
5. Start the transmitter. Adjust filament voltage to 27 volts.
6. Hold TEST KEY in non-lock position and adjust the FIRST-AMPLIFIER-TUNING and POWER-AMPLIFIER-TUNING dials until minimum plate current is indicated by M_6 and M_3 .
7. Adjust ANTENNA-TUNING dial so that maximum ANTENNA CURRENT is obtained as indicated by M_4 .

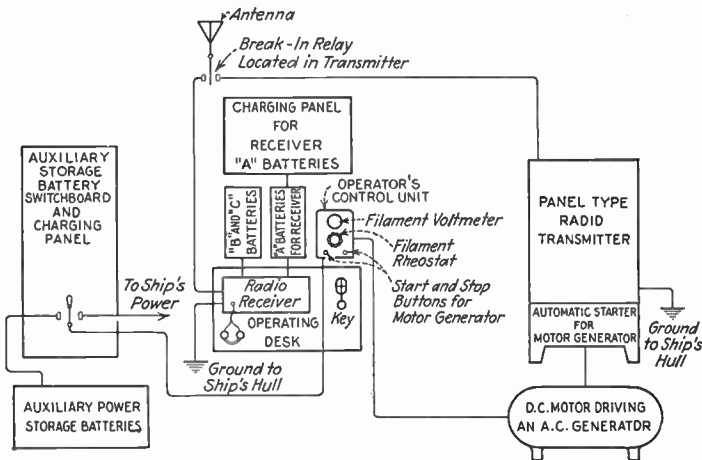


FIG. 273.—Block diagram of complete ship installation.

8. Readjust the POWER-AMPLIFIER-TUNING and ANTENNA-TUNING dials as before.
9. The POWER-AMPLIFIER PLATE current should be from 70 to 90 ma. (M_3). If too low, increase the ANTENNA COUPLING and retune. If too high, decrease it and retune.

10. Compare meter readings with typical limits given in table in instruction book accompanying the transmitter.

11. Set detents (catches) for this frequency.

12. For ICW transmission place CHANGE-OVER SWITCH in the ICW position. Hold TEST KEY down in non-lock position. The indication of the MODULATOR-PLATE current meter M_6 should be within specified limits.

13. For CW transmission place the CHANGE-OVER SWITCH in the CW position. Hold TEST KEY in the non-lock position and compare the meter readings with the typical limits specified in the table included in the instruction book.

Note 1: Another transmitter called model T6 is also supplied which is identical with T6A in all respects except that it is designed to operate from a primary source of power supplied by a 32-volt storage battery. Plate potential is supplied by a dynamotor which is operated by the storage battery.

Note 2: The adjustment of a transmitter to a desired frequency by the use of a heterodyne frequency meter is explained in Chap. XI on Radio Receivers, Sec. 290.

258. A Complete Ship-radio Station.—The layout of a complete ship station is shown in the block diagram (Fig. 273), in which the relation of one piece of equipment to another in the radio system is clearly shown. This arrangement is only general in nature, and the actual location of each piece of equipment depends upon the physical dimensions of the radio room. The arrangement here shown is, however, typical on American vessels, although somewhat different schemes are employed on many foreign ships.

Thus far, in this book, only the radio transmitter of the complete ship station shown in the block diagram has been treated. The remaining

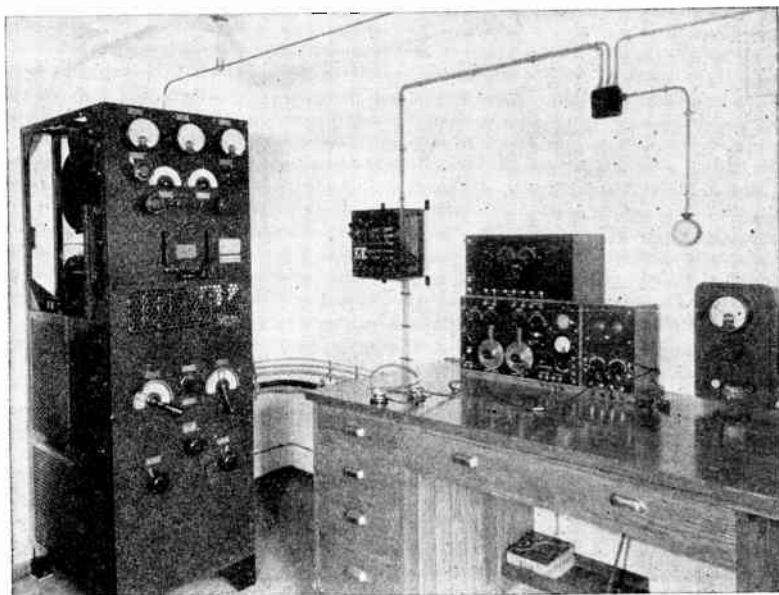


FIG. 274.—Modern installation on an American vessel. (Courtesy of R.M.C.A.)

parts will be treated subsequently, but, should the reader desire immediate information about any part shown, he has but to refer to the Index.

Modern radio direction finders are a part of the radio equipment of most of the larger vessels. This equipment is usually located in the pilot house or chart room adjacent to the navigating bridge and is, therefore, not shown in this block diagram. The radio operator is, however, usually responsible for the maintenance of the radio direction finder, although it is operated by the navigator himself or one of his assistants.

A photograph of a modern ship installation as illustrated previously by the block diagram is shown as Fig. 274, while a complete schematic wiring diagram may be had by joining Fig. 269 to 270.

259. Transmitter Care and Maintenance.—On a shipboard installation the transmitter may in time be covered in part by salt spray and

moisture which may allow high-voltage creepage along the panel or other insulation thus affected, resulting finally in a complete breakdown. To avoid such trouble the operator should go over all units and accessible parts daily with a clean dry rag.

All moving parts, except commutators and micarta bearings of tuning variometers, should be oiled to insure smooth operation. If radiation is low, examine the set for loose or broken connections. This may also cause excessive heating of the vacuum tubes. See that all grid leaks fit snugly in their clips.

All relays should be examined frequently to see that the contacts make and break properly. If the contacts are badly pitted, they may be filed down or replaced with new contacts.

See that all tubes are functioning properly, and that there is no blue flashing or change in filament temperature when the key is pressed. If this occurs new tubes may be tried throughout, but the old tubes should be immediately replaced if the new tubes do not remedy the trouble.

There should be no sparking about the set when the key is pressed.

Examine automatic motor starter monthly and see that contacts are kept clean and smooth; file down or replace when necessary.

Care and operation of the motor generator are completely discussed in Chap. XIV on Dynamo Electric Machinery.

If set is fitted with a chopper, keep motor oiled and wheel clean. Wipe commutator wheel with dry cloth daily.

Troubles, their causes and remedies are covered in Table VIII later in this chapter.

The contact surfaces of the rotary coils used in certain types of short-wave transmitters in the high-frequency tuning system should be cleaned by holding a piece of fine sandpaper against the outside surface and rotating the coil. Oil should never be used on the coil windings.

Antenna systems should be inspected weekly to make sure antenna wire or wires and lead-ins are held tightly in position. Clean deck and lead-in insulators weekly with a dry cloth. If antenna insulators become covered with carbon, and cleaning is ineffective, they must be replaced.

260. Transmitter Troubles.—The most satisfactory formula for preparedness against apparatus failure, in general, is to know your circuits. It is impractical to supply written directions for finding and curing all of the different troubles which may occur in a radio transmitter; the best that can be done is to list some of the most common troubles with their causes and remedies. Such a table follows, and the student or operator is urged to give it careful study.

When trouble occurs, go about finding the cause in a systematic manner. Inspect all the meter readings to localize the trouble which is usually in one portion of the apparatus such as the power supply, the

TABLE VIII.—TROUBLES IN VACUUM-TUBE TRANSMITTERS

Troubles	Probable causes	Probable remedies
1. Motor generator does not start	Open switch or fuse. Line voltage "off" or too low. If key relay works, line voltage is "on"	Close switches, or replace fuses. Close any open overload circuit breakers
2. Start contactor closes, but motor generator does not start	Burned-out starting resistor or open connection between starter and motor. Frozen bearing	Try shorting out burned-out sections of starting resistance. If fuses continue to blow, use water rheostat
3. Tube filaments do not light	Assuming generator is all right, look for blown fuse caused by filament shorted. Loose connection on, or defective, filament transformer, panel, or socket. Defective brush on motor slip rings. Defective tube. Open in filament rheostat	Test filament in each tube by continuity test with voltmeter. Open indicated by no reading; short indicated by full reading. Replace blown fuses or tighten loose connections. Replace brush if necessary. Repair rheostat or transformer (17, 18)
4. No radiation, tubes heat only to dull red (see also 23)	Antenna circuit out of tune. Grounded or open antenna circuit. Open or loose connections in r.-f. circuit in set	Tune set for maximum radiation. Shift to another wave band. Fix loose connections
5. Tubes heat badly when key is up	Improper blocking bias caused by defective resistor or open or loose connection. Soft amplifier tube may partly short bias voltage. Soft tube will probably flash blue and overheat	Fix connections or repair resistor. If a small section of resistor is burned out, defective part may be jumped with one or more 60-watt tungsten lamps. Replace defective tube
6. Tubes heat badly when key is pressed	Oscillator tube defective or oscillator circuit not oscillating because of loose connections. Defective grid leak	Remove all but oscillator tube. If tube gets too hot, circuit or tube is defective. Fix circuit or replace tube (see 19 and 20).
7. Amplifier tubes heat excessively when key is down and antenna circuit is in tune	Defective coupling transformer or insufficient number of plate turns in use	Fix coupling transformer or increase plate turns
8. Audio oscillator does not operate	Loose connections to switches, transformer, or condenser. Defective condenser or tube	Tighten all connections and replace tube
9. Extremely high plate current when key is pressed	Shorted grid leak somewhere possibly. Excessive voltage on filament or plate. Oscillator not operating	Fix or replace grid leak which is defective. Test with continuity test with voltmeter. Full reading indicates short circuit. Adjust voltages. Fix oscillator
10. Chopper motor does not start	Loose or open connections to motor. Poor brushes. No oil in bearings	Fix connections or close circuit. Replace brushes. Oil bearings
11. Insulation breakdown	Excessive potential. Salt deposit from atmosphere. Dirt on surface. Carbon on surface	Replace insulation if possible and necessary; clean off burned section; cut away burned section. Cut winding, and remove sufficient turns to clear; then jump removed section. If switches, remove them from the circuit and complete circuit with temporary jumper
12. Master-oscillator plate choke burned out	Defective choke, or too much current	Remove burned-out turns or replace by spare or by grid choke. Close grid circuit with jumper. Operate set on reduced power
13. Plate- or grid-blocking condenser shorting	Excessive current surges, or defective condenser	Remove condenser and substitute the tank condenser from another wave-band circuit or replace with an equivalent condenser from some other apparatus

TABLE VIII.—TROUBLES IN VACUUM-TUBE TRANSMITTERS.—(Continued)

Troubles	Probable causes	Probable remedies
14. Plate voltmeter not operating	Loose connections or defective meter	Maintain generator field rheostat at usual position, and watch tubes to see that they operate at normal temperature
15. Plate ammeter not operating	Loose connections or defective meter	Insert a 150-watt lamp in place of meter, and do not allow it to exceed full brilliancy when it replaces 0-5-amp. meter
16. Antenna ammeter inoperative	Overloaded or defective meter	Adjust set in accordance with tuning-record card, and observe plate ammeter for final resonance
17. Filament transformer burned out	Overloaded or defective transformer or circuit	Connect a sufficient number of heavy-duty storage cells in series to give required filament voltage and connect direct to filament bus bars with wire heavy enough to carry the filament-current load. Rheostat should be in circuit, but cut out if not necessary
18. Filament rheostat burned out	Short circuit in filament circuit; rheostat not heavy enough, or too old and worn out	Use lead cells as described above, but test total voltage with voltmeter. Do not exceed rated filament voltage of tubes. Cut rheostat out of the circuit
19. Grid leak open in master-oscillator circuit	Subject to overload or defective	The grid leak in an oscillator circuit may be replaced by a resistance of 4,000-10,000 ohms. A column of water, 12 in. long, in a rubber hose plugged at each end and having wire electrodes will be satisfactory. If the resistance is too high, add a little salt or washing soda
20. Grid leak open in power amplifier	Has been subjected to an overload or is defective	Substitute depends on number of tubes used, a 40-watt tungsten lamp equals 370 ohms; a 60-watt lamp equals 220 ohms. One or more lamps may be used in series, in parallel, or in series-parallel in an emergency depending on the tubes used
21. Tank condenser shorted or overheating	Overloaded or strained by excessive surges	Replace defective condenser by similar one taken from another tank circuit or by spare condenser
22. Failure to oscillate	<ol style="list-style-type: none"> 1. Fuses blown 2. High-potential plate supply circuit open (burned-out radio-frequency choke, poor plate connection on the tube, burned-out generator-filter condenser and choke, improperly fitted brushes on the generator commutator, broken leads) 3. Oscillatory circuits not adjusted properly. Improper values of plate and grid inductances or capacities 4. Burned-out grid leak or grid choke 	<ol style="list-style-type: none"> 2. Close plate supply 3. Tune oscillator 4. Replace defective leak or choke

TABLE VIII.—TROUBLES IN VACUUM-TUBE TRANSMITTERS.—(Continued)

Troubles	Probable causes	Probable remedies
	5. Plate- or grid-blocking condenser open or shorted (the latter may produce serious overheating of the tube) 6. Plate potential too low 7. Filament potential too low 8. Aerial or ground circuit open 9. Transmitter improperly tuned to resonance with the radiating system 10. Filament deactivated 11. Sagging filament touching the grid 12. Loose, dirty, or corroded connections, especially in the oscillatory circuit 13. Poor tube-socket connections	5. Replace defective condenser 6. Increase plate potential 7. Increase filament voltage 8. Inspect and repair antenna and ground 9. Tune amplifier and antenna system 10. Replace tube or reactivate by operating filament at normal temperature with plate power "off" for 10 min. or longer 11. Place tube in various positions, if possible, or replace tube 12. Fix poor connections 13. Tighten spring action in tube sockets and remove corrosion with fine sandpaper
23. No indication of radiation (see also 4)	1. Tube not oscillating 2. Defective antenna ammeter 3. Defective antenna insulation 4. Circuits not in resonance 5. Poor connections on the antenna 6. Improper coupling	1. See 22 2. See 16 3. See 11 4. Tune circuits to resonance 5. Solder all joints 6. Adjust coupling
24. Low radiation	Dirty or leaky insulators Circuits out of tune Improper coupling Improper voltages See causes and remedies under 23 some of which may cause low radiation. Defective crystal, oscillator tube, or interstage transformer	Clean or replace insulators Adjust circuits Adjust coupling Adjust voltages See 35; replace tube, or transformer
25. No plate voltage	Abnormally high screen-grid current Open relay; open door switch; open generator field rheostat; defective voltmeter; blown fuses; poor brushes on generator; no generator field excitation	Close relay and switches; repair rheostat; replace fuses after rectifying cause of blowout; clean commutator, or fix or replace brushes. Close excitation circuit on generator
26. No plate current	Open overload relay; poor plate-terminal connection on tubes; no voltage, see 25; defective or open plate ammeter; defective tube; grounded high-voltage line	Close relays; fix any loose connections; apply remedies given under 25; See 15; replace tube; clear high-voltage line of grounds, if circuit so demands
27. No grid current on oscillator	Plate voltage too low; poor connection on tube; poor keying contacts; open grid leak	Raise plate voltage; tighten all connections; adjust and clean keying contacts; replace grid leak
28. Too high grid current on oscillator	Disconnected or grounded antenna; antenna circuit out of tune; insufficient load on oscillator	Examine antenna and lead-in insulator, clean or repair; tune antenna circuit; tighten coupling to antenna

TABLE VIII.—TROUBLES IN VACUUM-TUBE TRANSMITTERS.—(Continued)

Troubles	Probable causes	Probable remedies
29. Keying relay inoperative	No line voltage; wave-change switches do not line up properly; contacts stuck together; open resistor in circuit; relay coil burned out or open	Close line circuit; adjust wave-change switches; clean contacts with fine sandpaper; replace resistor; repair magnet coil
30. Failure of chopper motor to start	No line voltage due to open switch or fuse; motor needs oil; commutator dirty, or brushes defective	Apply line voltage; oil motor; clean commutator and brushes; replace brushes
31. Ragged ICW signals	Dirty commutator on chopper; poor contact by brushes to chopper commutator	Clean commutator with No. 00 sandpaper. Fix brushes, and adjust pressure on commutator
32. Unsteady short-wave signals	Swinging antenna; loose high-frequency connections; too-close antenna coupling; erratic atmospheric conditions; wrong frequency being used	Fix any loose connections; adjust beyond control of operator if right frequency is being used
33. White deposit inside tube	Air leak; deposit shows up when voltage is applied to filament	Replace tube
34. Deviation of frequency	On crystal-controlled sets, failure of temperature-control equipment	Inspect thermostat; replace, if necessary
35. Failure of crystal oscillator to oscillate	Defective crystal caused by chipping, or dirt in holder; defective tube; open circuit	Clean crystal and electrodes using carbon tetrachloride and lens paper or other lintless material. Do not touch crystal surfaces after cleaning. Set upper electrode firmly on crystal; replace tube; look for open circuit
36. Too high r.-f. output	Antenna adjustment incorrect Incorrect or reversed screen voltage	Adjust antenna circuit and coupling Adjust screen voltage to proper value
37. One parallel tube heats badly	In amplifier circuits using two tubes in parallel, the trouble is due to the tubes not dividing the load	Check wiring or interchange tubes until two are found which will divide the load

rectifier or motor generator, the oscillator, the amplifier or the antenna system. Compare present performance with normal performance; this should tell the operator a great deal.

Continuity tests can be made with a high-resistance voltmeter and, in many instances, a battery. On speech-amplifying equipment, output can be checked by listening across the secondary output impedance of the output transformer with a pair of high-resistance headphones in series with a blocking condenser of 1 or 2 mf. capacity.

If C batteries are used, they should be checked occasionally and replaced when their voltage has fallen 15 per cent.

Radio-frequency output voltage can be checked by means of a small neon light. In high-power stages, the presence of radio-frequency energy can be determined by touching the high-potential end of the tank with a small coil of two or three turns of bare copper wire on the end of an insulating stick or handle. Radio-frequency potentials of 200 to 300 volts or more will cause arcing to the above mentioned coil.

When other remedies fail it is always advisable to change tubes as spurious conditions often exist in an apparently good tube.

261. Coastal Stations.—Transmitters in coastal stations are for the most part especially built for the particular service they are expected to perform. For some communication purposes conventional ship transmitters operated from three-phase power lines are used, while for other purposes a special model is built for the work to be performed. Coastal stations are called on to communicate over very long distances by short-wave and long-wave emissions, and over short distances (around 500 miles or less) by intermediate-wave emissions. For long-distance work, *CW* is used exclusively; whereas for short-distance work, both *CW* and *ICW* are used.

Long-distance communication to vessels in the north and south Atlantic Ocean is handled by the Radiomarine Corporation by their stations at Chatham, Mass. (WCC), and Tuckerton, N. J. (WSC), and by the Mackay System by their high-powered station at Sayville, L. I., N. Y. (WSL). Traffic over long distances in the Pacific Ocean is handled by the R.M.C.A. station at Bolinas, Calif. (KPH), and by the Mackay System station near Palo Alto, Calif. (KFS). Numerous other stations are provided on both coasts for short-distance communication.

The transmitting equipment at a typical coastal station is that provided at Tuckerton, N. J. (WSC), which is tabulated below, eight transmitters being provided.

Approximate wave length, meters	Frequency, kilocycles	Power
17.79	16,860	1 kw.
23.66	12,675	40 kw.
26.84	11,175	200 watts
35.60	8,430	40 kw.
47.24	6,350	200 watts
47.31	6,340	1 kw.
600 and 649.3	500 and 462	5 kw.
2,255 and 2,400	133 and 125	3 kw.

In order that duplex operation may take place, the transmitters at high-powered stations are located at some distance from the receiving station. For example, at Chatham, Mass. (WCC), the transmitters are located at Marian, Mass., a distance of 50 miles, whereas at Tuckerton, N. J. (WSC), the transmitters are located 5 miles from the receiving station. In this way, during periods of heavy traffic, a dozen or more ships can be worked simultaneously.

At lower powered coastal stations, as for example WNY, the New York Harbor station, both transmitters and receivers are located under one roof and are generally of the type found aboard ship.

II. LOW-POWER RADIO TELEPHONE AND TELEGRAPH TRANSMITTERS

262. Radio-telephone Transmitter, 400 WATT: W.E. 9C. *a. Description.*—The complete wiring diagram of this transmitter together with its power supply, the 2B rectifier, is shown in Fig. 275 A and B. This transmitter is ideal wherever a low-power unit is desired. It is designed for radio-telephone transmission on frequencies between 1,500 and 6,000 kc./s. with a carrier output of approximately 400 watts. It is capable of substantially complete modulation, and its frequency range and power make this transmitter very desirable for aeronautical or police requirements.

The circuit employs a crystal-controlled oscillator system using two separate quartz crystals for operation on any two frequencies within its frequency range as required.

b. Theoretical Operation.—With the proper adjustment of the load impedance L_1 for the crystal unit Y_1 or Y_2 , a voltage will be developed through the inter-electrode capacity between G and P of the oscillator tube V_1 . This excites the crystal through the establishment of a voltage across it which starts it oscillating. The oscillating voltage developed across the crystal is applied across GF of the oscillator tube which produces radio-frequency oscillations across the load impedance L_1 at the frequency period of the quartz crystal. These oscillations are then applied across GF of the amplifier tube V_2 . If the transmitter frequency is to be between 1,500 and 3,000 kc./s., the plate-load impedance of V_2 is tuned to the frequency developed by the crystal, but if the desired frequency is to be between 3,000 and 6,000 kc./s., then the plate-load impedance is adjusted to resonance at a frequency twice that of the crystal. In the former case V_2 is functioning merely as a first radio-frequency amplifier, while in the latter case the tube is used as a radio-frequency doubler. The radio-frequency voltages developed across the load impedance of V_2 are then applied to the grid of the second radio-frequency amplifier V_3 and finally to the final or power stage V_4 .

The Rice method of neutralization is used to stabilize each amplifier circuit. The second amplifier V_3 also functions as the modulating radio-frequency amplifier tube when audio-frequency variations are applied to it. When a microphone input is applied to the primary winding of T_3 , voltage variations are applied to the three paralleled grids of the speech-amplifier tubes, V_5 , V_6 , V_7 , by induction into the secondary of T_2 . These variations are then amplified in the plate circuit and pass through the modulation transformer T_1 . The audio-frequency voltages, then developed across its secondary, modulate the radio frequencies developed in the load circuit L_2 in accordance with the Heising principle of modulation.

These voltages are then passed on to the power amplifier to be further amplified and finally radiated from the antenna.

c. Changing Frequency.—In the aviation service the band from 2,900 to 3,500 kc./s. is used at night, and the band from 4,900 to 5,700 kc./s. is used during the daytime. Assuming that the night frequency used is 2,922 kc./s. (102 meters), and the day frequency is 5,612.5 kc./s. (53.4 meters), the procedure for changing to either of these frequencies is as follows.

1. Set the crystal grid connection switch to the proper unit. The day crystal unit is small and round, the *night* crystal large and square. Set output transfer switch D_7 to dummy-antenna load.

THEN MAKE THE FOLLOWING SWITCH ADJUSTMENTS:

2. Set oscillator plate-impedance switch D_1 to tap.....	Night 7	Day 6
3. Set intermediate-amplifier coil switches D_2 and D_3 to.....	Low	High
4. Fix position of intermediate-amplifier link D_{14}	Closed	Open
5. Set intermediate-amplifier resistance switch D_{15}	Open	Closed
6. Set modulating-amplifier tuning-coil switch D_9 to turns.....	16	8
7. Set power-amplifier plate-coil switch D_5 to turns.....	7	3
8. Fix position of power-amplifier plate-condenser switches D_4 and D_6 to.....	Closed	Open
9. Switch to proper antenna		

THEN TUNE TRANSMITTER AS FOLLOWS:

10. Push START button on front of 2B rectifier panel, and, when time-delay relay operates, close microphone-stand THUMB SWITCH and make the following adjustments on tuning dials using data on typical meter readings supplied with transmitter as guides.

11. Open the plate circuits of the modulating-amplifier tube and power-amplifier tube by opening D_{11} and D_{12} . Adjust the INTERMEDIATE-AMPLIFIER PLATE-TUNING dial to obtain a maximum MODULATING AMPLIFIER GRID CURRENT.

12. Adjust the MODULATING-AMPLIFIER PLATE-TUNING dial to obtain a minimum of the rectified MODULATING-AMPLIFIER PLATE CURRENT. This adjustment requires that the switch blade D_{12} touch the frame in order to complete the circuit from plate to ground.

13. Close the modulating-amplifier plate circuit by closing D_{12} . Turn the MODULATING-AMPLIFIER PLATE-TUNING dial to obtain a minimum MODULATING-AMPLIFIER PLATE CURRENT.

14. Close the output transfer switch D_7 to the load-resistance side, and close the power-amplifier plate switch D_{11} . Increase the POWER-AMPLIFIER GRID-INPUT dial until the POWER-AMPLIFIER CURRENT is about 400 ma. Vary the POWER-AMPLIFIER PLATE-TUNING dial to obtain minimum plate current. Gradually increase the POWER-AMPLIFIER GRID-INPUT adjustment, keeping the plate circuits of the modulating-amplifier tube and the power-amplifier tube tuned, until the desired output is obtained. The output power should not exceed 400 watts or 0.9 amp. into the 500-ohm load resistance R_{12} .

15. Note that all meter indications are near those given in the table of typical meter readings which are contained in the instruction manual accompanying the set. After one or two shifts the operator should have these readings for the DAY or NIGHT frequency pretty well in memory.

16. To check the neutralization of the modulating amplifier shut down the transmitter and remove the connection D_{13} from the top electrode of the quartz-crystal frequency control, and connect it to the crystal-heater supply wire or any other good ground. Start the transmitter and note whether or not the OUTPUT RADIO-FREQUENCY-CURRENT and the MODULATING-AMPLIFIER GRID-CURRENT meters indicate. If these meters indicate, the neutralization is not satisfactory. If they do not indicate, the transmitter is ready to operate when the connection D_{13} is replaced on the frequency control, the transmitter being shut down for this replacement, and the output-transfer switch D_7 thrown from the dummy load to the antenna.

d. Maintenance and Trouble Hunting.—The maintenance of this transmitter is limited to keeping the set clean, seeing to it that all relays work properly, and anticipating tube troubles. As far as possible, tube

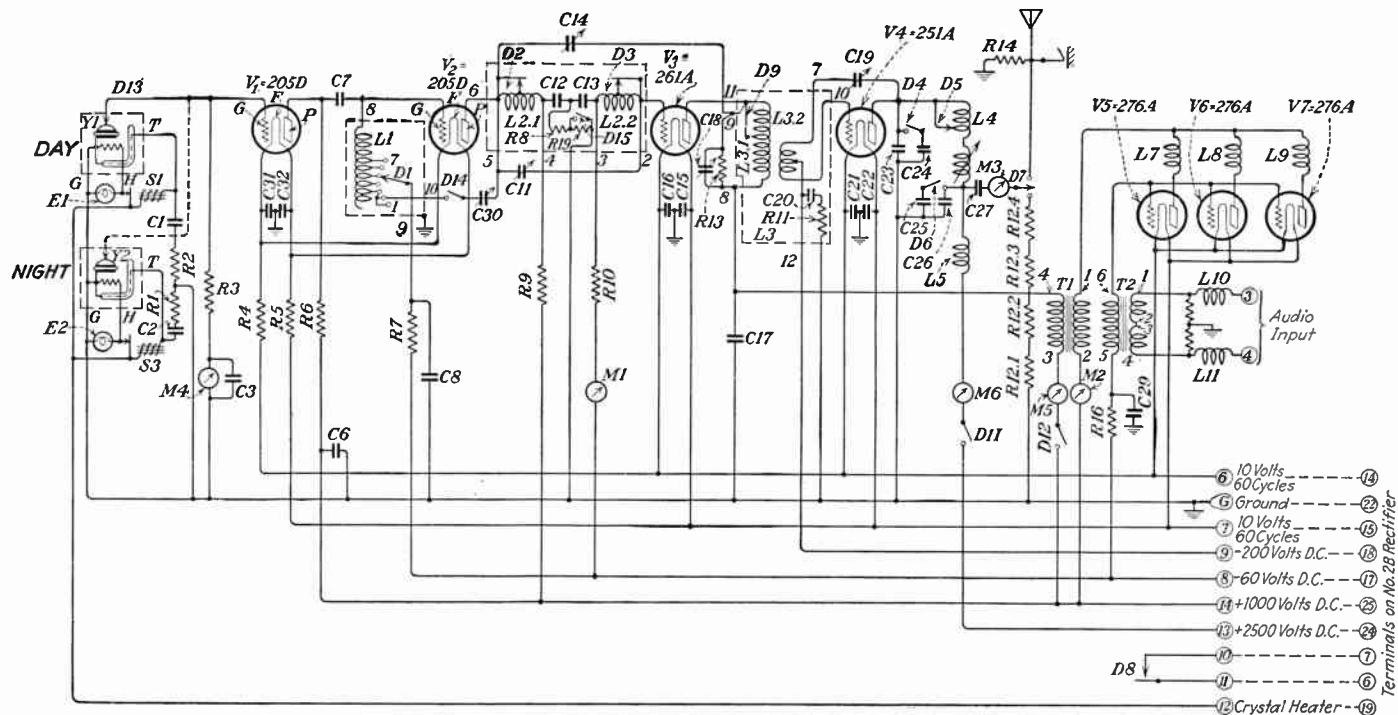
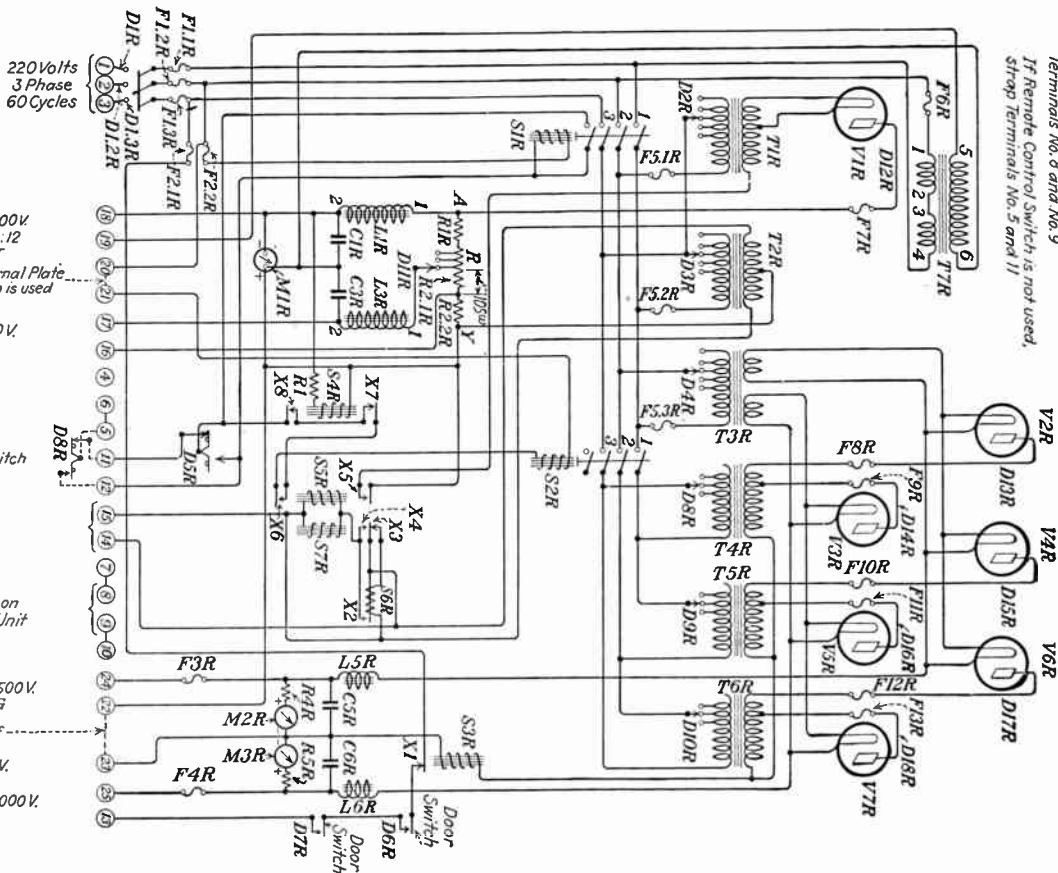


FIG. 275A.—Simplified schematic diagram of low-power radio-telephone transmitter; Western Electric 9C.

If No. 3-4 Tuning Unit is not used, strap
Terminals No. 8 and No. 9
If Remote Control Switch is not used,
strap Terminals No. 5 and 11



To Terminal No. 9
on Transmitter -200V
To Terminal No.12
on Transmitter

Remove Strap if External Plate
Voltage Control Switch is used

To Terminal No.8
on Transmitter -60V
To Terminal No.2
on Transmitter

To Terminal No.11
on Transmitter

Remote Control Switch

To Terminal No.7
on Transmitter
10 V.A.C.

To Terminal No.6
on Transmitter
To Terminal No.10
on Transmitter
To Gate Switch on
No. 3-A Tuning Unit

To Terminal No.13
To Terminal No.13
on Transmitter +2500V.
Ground to Terminal G
on Transmitter

Remove Strap if
Switch is used
-2500V. and -1000V.

To Terminal No.14
on Transmitter +1000V.
To Terminal No.10

FIG. 275B.—Low-power rectifier unit; Western Electric 2B. (Courtesy of Western Electric Company.)

failures should be anticipated and the proper tube replacements made beforehand. Keeping a record of the length of time a tube operates, and observing the conditions of the elements inside the tube, are of help in guarding against tube failures. The possibility of internal tube short circuits is increased when the plate or grids become warped or when the filament sags. All tubes should be tested frequently for normal plate current, activity, and other characteristics.

e. Troubles in this transmitter are usually due to improper operation of some of the switches or adjustments, and the first thing to do in case of trouble is to check up on these.

Failure of radio-frequency output may be due to any number of things, and a typical procedure for locating the trouble would be to check first the filament current of all tubes with high voltage "off"; then proceeding to check each section of the circuit beginning with the oscillator. Check position of *all* switches and links for the particular frequency being used.

Check the oscillator: see that the crystal is oscillating by noting oscillator grid current and that correct crystal is being used; check operating temperature by noting that indicator lamp flashes slowly; try operating oscillator on another frequency; look for loose leads. When oscillator is checked as O.K., see that meter readings conform to proper operating values. Do this with each unit as it is checked O.K.

Check the modulating-amplifier grid current; if there is no grid reading, check tuning of intermediate-amplifier circuits, making sure this stage is not tuned to an undesirable harmonic of the oscillator. If these tests do not remove the trouble, the wiring should be checked for open or short circuits. Make sure that all switches are making good contact.

If the power amplifier is operating properly on the load resistance (dummy antenna), but there is no power in the antenna and the power-amplifier plate current is high, then the trouble is usually that the antenna is not tuned properly or that the transmission line is open.

When the radio-frequency output is low, the cause may be low filament voltage, poor contacts at tube sockets, improper tuning of some circuit, or poor tubes. If the trouble persists, try replacing the tubes one at a time with a new tube to make sure that a poor tube is not the cause.

Improper modulation is the trouble if the output radio-frequency current does not rise 10 to 20 per cent during passages of loud sustained tone.

If the speech amplifier is not operating properly, it will show up as failure to modulate. Check the leads from the speech-input equipment for open connections using a battery and a pair of phones. With the audio-input circuit correct as far as the terminal board, speak into the microphone and notice the fluctuations of the audio-amplifier plate current. It should be possible to vary this about 20 per cent with heavy

CERTAIN PARTS OF W.E. 9C TRANSMITTER (See Fig. 275A)

Condensers			Resistors		
Item	Microfarad	D.-c. volts	Item	Ohms	Per cent, + or -
C_1 }	1.0		R_1 }	150	10
C_2 }					
C_3	0.008		R_3	100,000	
C_6	0.01	2,000	R_4 }	0.07	5
			R_5 }		
C_7	0.002	2,000	R_6	14,000	10
C_8	0.01	500	R_7	12,000	
C_{12}	0.01	2,000	R_8^*	100	
C_{13}	0.01	700	R_9	16,000	5
C_{15} }	0.008		R_{10}	12,000	10
C_{16} }					
C_{17}	0.001		R_{11}^*	4	
C_{20}	0.01	500	R_{12}^\dagger	120	
C_{21} }	0.01	100	R_{13}	7,000	10
C_{22} }					
C_{23} }	0.00015		R_{14}	20,000	10
C_{24} }					
C_{25} }	0.0003		R_{16}	100,000	
C_{26} }					
C_{27}	0.01		R_{18}	600 CT	
C_{29}	1.0		R_{19}^*	400	
C_{31} }	0.008				
C_{32} }					

Meters

M_1 Grid milliammeter, 0-50 M_4 Grid milliammeter, 0-5
 M_2 Plate milliammeter, 0-500 M_5 Plate milliammeter, 0-200
 M_7 Radio-frequency ammeter, 0-1.5 M_6 Plate milliammeter, 0-1,000

* Ward Leonard Vitrohm Plaque, non-inductive, used as anti-sing resistor.
 † Load resistor, four 120-volt carbon-filament lamps, 120 watts.

CERTAIN PARTS OF W.E. 2B RECTIFIER (See Fig. 275B)

Condensers				Resistors		
Item	Number paralleled	Microfarad, each	Volts	Item	Ohms	Per cent + or -
C_1R	6	1.25	200	R_1R	440	5
C_3R	3	4.0	55	$R_{2,1}R$	40*	5
C_5R	2	0.85	2,500	$R_{2,2}R$	340	10
C_6R	4	0.85	1,000	$R_{2,3}R$	150	5
				R_4R	Weston multiplier	
				R_5R	Weston multiplier	
Fuses				Filter Inductances		
Item	Amperes	Volts	Item	Minimum henrys	Maximum ohms	Volts
F_1R	20	250	L_1R	30	285	200
F_2R	1	250	L_3R	30	285	55
F_3R	1	2,500	L_5R	7.3	36	2,500
F_4R	1	2,500	L_6R	4	27.3	1,000
F_5R	2	250				
F_6R	$\frac{1}{4}$	250				
F_7R	1	2,500				
to $F_{13}R$						

Meters

M_{1R} Grid bias voltmeter, 0-300 direct current
 M_{2R} Rectifier voltmeter, 0-3,000 direct current
 M_{3R} Rectifier voltmeter, 0-2,000, direct current

* In steps of 10, 10, 10, 10 ohms.

modulation. A much greater percentage increase in speech-amplifier plate current than this usually indicates a faulty modulation transformer. The modulating-amplifier grid current must be at least 8 ma. in order to obtain complete modulation. When modulation trouble occurs it is also well to watch the speech-amplifier tubes for any weakness in the tubes themselves which may generally be identified by the overheating of one of the tubes. Check the tubes by placing them, one at a time, in one of the audio-amplifier sockets and noting whether the plate current taken by each tube is approximately the same. Faulty tubes should be replaced as soon as discovered. Another cause of poor modulation may be low filament voltage on the power-amplifier tube. This will cause low filament emission and reduce the peak power output from the tube. Still another cause may be open-circuited filter condensers in the power-supply unit.

Many other causes of trouble are listed in the Table of Troubles (Table VIII) appearing earlier in this chapter.

263. Low-power Rectifier Unit.—The 2B rectifier, a Western Electric product, was designed to supply filament, plate and crystal-heater power, and the grid-bias voltages for operating transmitters 9A, B, or C or any other similar transmitter. The schematic circuit diagram is shown in Fig. 275(B).

a. Operation.—The rectifier unit is put into operation by depressing the starter button D_5R which energizes the solenoid of the relay unit S_1R from one side of the 220-volt alternating-current supply line, through the contacts X_1 of the relay S_3R and returning to the other side of the alternating-current supply line. A remote control switch may also be provided to effect the same starting procedure from a distant operating point, should it be desired to place the rectifier unit at a point removed from the transmitter.

If the button D_5R is pressed, the contactor arm will be attracted to the solenoid relay S_1R which closes switches 1, and 2, and 3. This action energizes the primary windings of the single-phase power transformers T_1R , T_2R , and T_3R . These transformers supply the filament potentials of 10 volts to the main rectifier-tube filaments, transmitter-tube filaments, and to the grid-bias rectifier-tube filament V_1R . Connected across the filament-voltage line is a thermal-controlled or time-delay relay S_6R which heats up as soon as the filament voltage is applied. The element gradually expands due to the heat and, after about 15 sec., closes the contacts X_2 . When these contacts are closed, the filament voltage energizes the two solenoids S_5R and S_7R . When these solenoids are energized, the contacts X_4 , X_6 , and X_5 are closed while the contacts X_3 open and disconnect the heating element of the relay S_6R . The closing of contacts X_4 completes the circuit to the plate of the grid-bias rectifier tube V_1R which applies the proper value of grid

bias to the transmitting tubes. The contacts X_6 are also closed simultaneously with X_5 since both are operated by the S_5R solenoid. It will be noted that the closing of X_6 does not yet energize the solenoid of relay S_2R since two other contacts X_7 and X_8 must also be closed because the three contacts are connected in series. However, at the instant the bias rectifier tube V_1R receives its plate potential by the closing of the relay contacts X_5 , a direct current passes through the voltage-divider resistance R and develops a potential of 200 volts, between the points A and Y . Hence, since the solenoid of the relay S_4R is connected across these points through the filter choke L_1R and a protective resistance R_1 , it will be energized. This action attracts the contactors, and closes X_7 and X_8 . Thus, when X_6 , X_7 , and X_8 are closed, the solenoid of the high-voltage relay S_2R will be energized which attracts the contactor and closes the switches 1, 2, and 3 located directly above it. This completes the 220-volt alternating-current supply to the three-phase rectifier system by energizing three primary windings of high-voltage transformers T_4R , T_5R , and T_6R . These transformers develop two separate high voltages, one of 1,000 volts for the plate supply to the speech amplifier, crystal, and first and second radio-frequency amplifier tubes; and 2,500 volts for the final or power-amplifier stage. The 1,000-volt positive lead is taken from the filaments of the rectifier tubes V_3R , V_5R , and V_7R , and the 2,500-volt positive lead, from the filaments of the V_2R , V_4R , and V_6R rectifier tubes. All of these tubes are of the mercury-vapor type.

b. Maintenance.—The following points are important in connection with the operation and maintenance of this rectifier unit.

1. After long operation, relay contacts may need cleaning.
2. Check relay S_7R to see that one contact is broken just after the other.
3. For noisy relays, adjust by bending armature slightly with a pair of pliers so that, when in the operated condition, the armature rests evenly against the pole pieces.
4. Watch mercury-vapor tubes to see that they have a "healthy" blue glow. Replace tubes as soon as possible when blue glow becomes dull which indicates life of tube will be short. If a mercury-vapor tube becomes inoperative in a high-voltage circuit, it may perhaps be used in a low-voltage circuit. Start new tubes in grid-bias rectifier, and replace poor tubes in high-voltage circuit with one from next lowest voltage circuit.
5. Allow filaments of new tubes to burn for 20 to 30 min. before applying anode voltage. This may be done by removing anode clip and starting rectifier.

264. Converting a Phone Transmitter for CW.—In order to use a radio-telephone transmitter for telegraphy, it is necessary to convert it for *CW* or *ICW*. Such a conversion is simply a problem of keying the transmitter so that the carrier wave may be interrupted. There are a number of methods of doing this and the particular one to adapt would depend on the power of the transmitter and the equipment available.

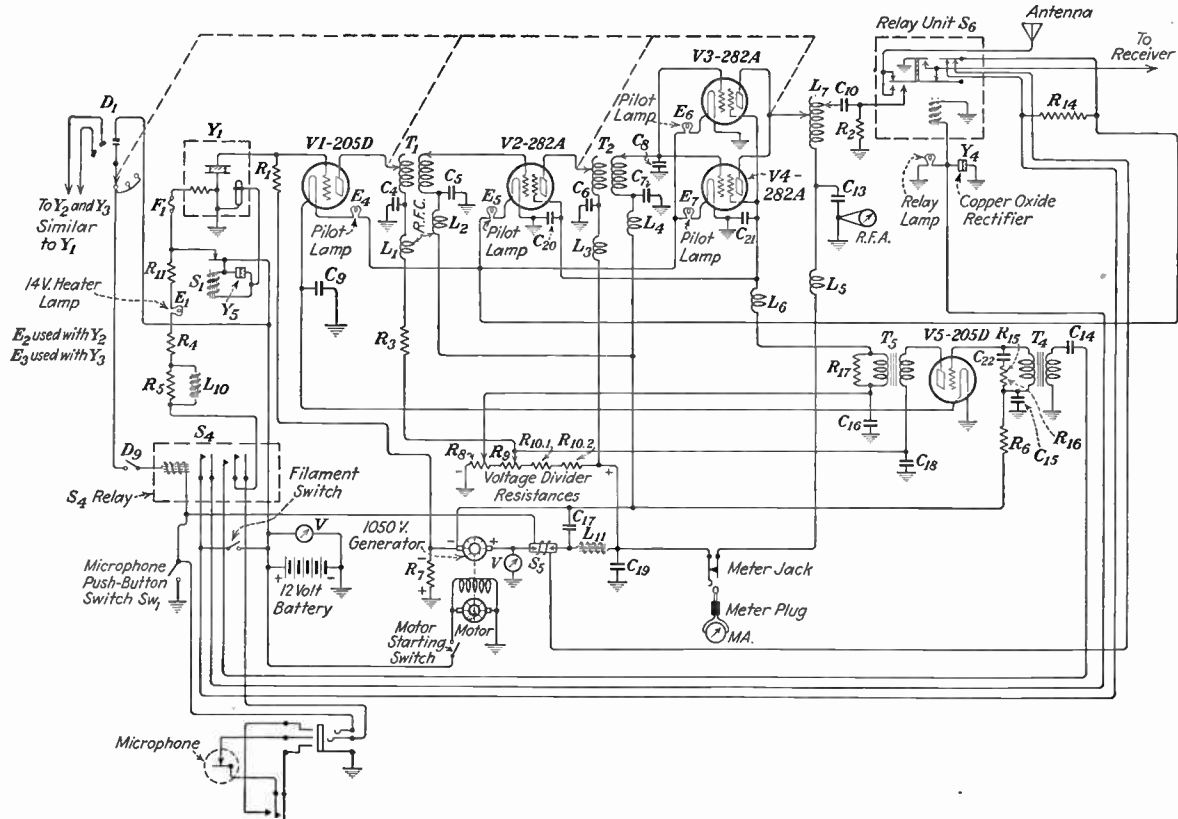


FIG. 276.—Simplified diagram of aircraft radio-telephone transmitter; Western Electric 13A.

CONDENSER AND RESISTANCE VALUES: W.E. 13A TRANSMITTER (See Fig. 276)

Condensers				Resistances		
Item	Farads	± Per cent	D.-c. volts	Item	Ohms	± Per cent
C_4 } to } C_7 }	0.004 mf.	20	3,500	R_1 } R_6 } R_{16} }	100,000	
C_8	75 mmf.	5		R_3^*	2,000,000	5
C_9 } C_{22} }	0.01 mf.	20	700	R_4	2,000	
C_{10}	4,000 mmf., 250 mmf., 125 mmf., or 62½ mmf. as required	5		R_5	100	
C_{13}	0.004 mf.	5		R_7 } R_{11} } R_{12} } R_{13} }	65	5
C_{14} }				R_8^*	1,200	5
C_{16} }				R_9	5,000	2
C_{17} }	1.0 mf.			$R_{10,1}$ }		
C_{18} }				$R_{10,2}$ }	2,350	2
C_{15}	0.25 mf.			R_{14}	0.625	20
C_{19}	3.0 mf.			R_{15}	500,000	
C_{20} }				R_{17}	5,000	
C_{21} }	0.004 mf.	20	3,500			

* To be wound with wire not smaller than No. 40 (0.0031 in.).

To convert a low-power transmitter, not over 500 watts, for *CW*, the simple method of interrupting the negative high-voltage return to filament on the power amplifier at the point of lowest potential, which would be close to the ground connection, would be satisfactory. Such a scheme is shown in Fig. 261(c).

If a power higher than 500 watts is involved, the conversion is not so simple. However, the simplest way of keying such a transmitter would be to insert a resistor in series with the negative high-voltage return to filament of such a value that the voltage developed across it would be sufficient to bias the amplifier tubes to cut-off. The key could then be arranged to short-circuit this resistance when keying as shown in Fig. 261(b).

A method of producing an audio-frequency output in the carrier is to fix a buzzer in front of the microphone and operate the transmitter in the regular way, keying being accomplished by inserting a key in series with the buzzer.

265. Aircraft Radio-telephone Transmitter: W.E. 13A. *a. Description.*—This radio-telephone transmitter is designed for transport plane service and is of the crystal-controlled type. It is capable of substantially complete modulation and of delivering a power output of 50 watts on three different (selected) frequency channels, any one of which may be selected by a single mechanical control which may be remotely operated from any point in the plane.

The transmitter, a schematic diagram of which is shown in Fig. 276, operates from a direct-current supply furnished by a small dynamotor which is capable of delivering the required e.m.f. of 1,050 volts at a current load of 350 ma. The filaments of the transmitter are lighted by a 12-volt storage battery. This battery also drives the dynamotor for the high-voltage plate supply. All circuits are well protected by relays and fuses and provided with the necessary controls to effect simple and efficient operation.

A typical arrangement of a transmitter in an airplane is shown in Fig. 277. The exact placement of each part will, of course, vary with the type of equipment used and the space available in the ship. Great care must be exercised to achieve proper weight distribution so as not to interfere with the stability of the plane.

An important rule to be remembered by aviation radio men is: *Never operate the radio transmitter while the gas tank is being filled.*

b. Practical Operation.—Assuming that the dynamotor is in operation, the closing of switch D_2 energizes the solenoid of the relay S_4 . The frequency-changing switch D_1 must also be in the operating position for any one of the three available frequencies, otherwise no high voltage will be applied to the plates of the tubes. When the microphone switch Sw_1 is pressed, two sets of contacts on the control relay S_4 close. One

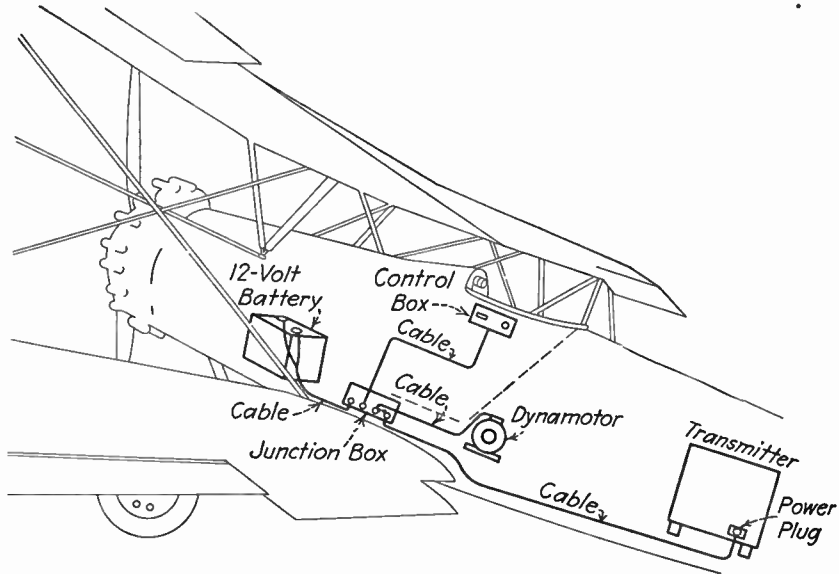


FIG. 277.—Typical radio installation in an airplane.

of these energizes the solenoid of the antenna-relay unit S_6 . This same contact may be used to start the dynamotor, since it controls the positive lead of the 12-volt battery. If, therefore, a connection is made to the proper terminal, simultaneous operation may be obtained the moment the microphone push-button is closed. In the diagram of Fig. 276 a separate switch is shown in the dynamotor-starting circuit connecting to the positive battery lead for simplicity reasons only.

The other contacts on relay S_4 , which close when switch Sw_1 is pressed, close the microphone circuit to the speech-input equipment. This feeds the audio-frequency voltages developed by the microphone into the primary winding of the microphone transformer T_4 . The energizing of the antenna-relay solenoid also closes a vacuum relay S_5 , which in turn closes a pair of contacts and applies the positive potential of the high-voltage source to the voltage divider and the plates of the vacuum tubes.

The heater circuits of the three crystal-controlled units are controlled by three relays S_1 , S_2 , and S_3 . These units operate in conjunction with a thermostat similar to that previously explained in Sec. 249 on the broadcast transmitter. Three copper oxide rectifiers Y_6 , Y_6 , and Y_7 are provided merely for the purpose of preventing sparking at the thermostat contacts. Commutator-ripple variations are smoothed out by means of the filtering system including L_{11} , C_{17} , and C_{19} . This enables a steady direct current to be applied to the plates of the vacuum tubes for quiet operation. The biasing potential for operating tubes V_2 , V_3 , V_4 , and V_5 at the proper portion of their operating curves is provided by the voltage drop across the resistance R_7 . In other words, since all of the grid-return leads of these tubes terminate at the negative terminal of the high-voltage generator, and the filaments are at ground potential, the point $-$ on the biasing resistor R_7 is more negative than that of the grounded point $+$.

c. Theoretical Operation.—When the proper plate, bias, and filament voltages are applied to all of the tubes, a plate current will flow in all of the primary windings

of the radio-frequency transformers T_1 , T_2 , and L_7 and through the audio-frequency transformer T_5 , establishing a potential drop across them. Owing to the fact, however, that the primary winding of T_1 is adjusted to resonance with the crystal frequency (assuming the proper crystal and coils are used), the voltage developed across its primary winding will be applied through the inter-electrode capacity of the tube V_1 across the crystal in the grid circuit. This potential sets the crystal into oscillation at a frequency depending upon the thickness and temperature of the crystal and the temperature-control chamber. These oscillating voltages will then act and react between the plate and grid circuits, in accordance with the principles of oscillation, at a definite frequency depending upon the constants of the circuit. In this particular transmitter the output circuit of T_1 and the plate circuit of T_2 are tuned to the second-harmonic frequency of the crystal oscillator. Hence, the frequency of the voltage variations developed in the plate circuit of T_2 are not of the fundamental frequency of the transmitter but a frequency of twice that value, and the V_2 -tube circuit constitutes a frequency-doubling system. This method provides greater circuit stability, since the crystal itself may be of a larger physical structure and it also provides less reactions with the oscillator circuit owing to possible load changes in the amplifier tube circuits.

The second-harmonic frequency is then amplified through the two power-amplifier tubes V_3 and V_4 , provided that the plate-load inductances L_7 is adjusted to the harmonic frequency. The radio-frequency variations now developed across L_7 may be caused to excite the antenna radiating system by the proper choice of the antenna series-condenser C_{10} and the adjustment of the movable clip on the antenna coil L_7 . This condenser should be changed and the clip varied until a maximum deflection is obtained on the antenna radio-frequency ammeter. This maximum deflection will be accompanied by a decrease in the plate-current reading of the second-amplifier tubes V_3 , V_4 , as indicated by the meter MA. If a steady tone, such as produced by a high-frequency buzzer or audio oscillator, is then directed into the microphone, the low-frequency variations in the microphone circuit will be fed through the transformer T_4 , and to the grid of the speech-amplifier tube V_5 . These low-frequency voltage variations, developed across the grid filament of V_5 , will cause corresponding frequency variations in the primary winding of the modulation transformer T_5 . These variations in turn will develop an e.m.f. across its secondary winding. Now since the screen-grid voltage of the amplifier tubes is supplied through this winding, any variation in voltage across it will affect the stability of the screen voltage. Hence, if this voltage is varied at an audible rate by the existing audio-frequency voltages across the secondary of T_5 , the plate circuit of the amplifier tubes will be varied in accordance with it. The variation of the radio-frequency current in the plate circuit of the amplifier tubes by this action is the modulation process in which the voice frequencies are transmitted into the ether. In order to insure that this condition is fulfilled, the buzzer or audio oscillator is used to provide a steady single tone and its effect upon the radio-frequency antenna current noted. The meter (RFA) should show an increase in antenna current over the unmodulated condition of approximately $22\frac{1}{2}$ per cent or no less than 20 per cent. This rise in antenna current is accompanied by a corresponding increase in the amplifier plate current as shown by meter MA of approximately 15 per cent. If the reading on MA during modulation does not increase more than 10 per cent, or show a drop below the normal unmodulated-current readings, the inductance in the plate circuit must be decreased by moving the adjustable clip one turn lower on the plate side of the coil L_7 . The antenna circuit, must, however, then be retuned to insure maximum conditions of resonance. If, on the other hand, the modulation is excessive or more than $22\frac{1}{2}$ per cent, the inductance of L_7 must be increased and the antenna retuned.

(d). *Maintenance.*—Where a number of equipments must be serviced, it is advisable to install a test bench such as is shown in Fig. 278. By its use all the transmitters can be kept in good condition, failures can be anticipated, and interruptions in service avoided.

Keep the apparatus clean. A cloth moistened with chemically pure carbon tetrachloride will be useful in cleaning exposed insulators. Occasionally inspect nuts and screws, and tighten if necessary.

Adjust any improperly working relays. The contacts of the control and heater relays may be cleaned with the flat end of a toothpick dipped in chemically pure carbon tetrachloride and polished with a special burnishing tool supplied by the manufacturer. Abrasive cloth or paper should not be used.

e. Location of Trouble.—If no radio-frequency output is obtained, make sure filament, bias, and plate voltages are applied at correct values. Measure oscillator grid and plate currents. No plate current indicates an open circuit which may be located by referring to the diagram and testing with an ohmmeter. No grid current indicates the quartz plate is not oscillating, which in turn may be due to a defective quartz plate or oscillator tube or an open circuit. Replace the tube and quartz plate, and check the circuit with an ohmmeter. If the oscillator is working properly, check the first amplifier and then the second amplifier by a similar procedure.

If the radio-frequency output is below normal, measure the screen voltage using a voltmeter of not less than 1,000 ohms per volt. If the voltage is low, one or more of the first- or second-amplifier tubes are probably drawing an abnormally large screen current. This may best be done by substituting three new tubes and by replacing the new ones with an old one, one at a time, until the poor tube has been found. If it is a second-amplifier tube and is not in too bad condition it may give additional service in the first-amplifier stage.

Other possible causes of low output may be a defective quartz plate, oscillator tube, or inter-stage transformer.

If the output is normal but there is no modulation, an open circuit or defective apparatus in the audio amplifier is indicated. It may be checked as follows: Plug a low-range milliammeter into the audio grid jack and make a loud sustained sound in the microphone. If the input circuit and transformer are correct, it should be possible to draw a fraction of a milliampere grid current on the audio tube. The audio-amplifier plate current should be within normal limits if the tube and circuit are correct. It should change little in either direction when the operator speaks into the microphone. A large change indicates a defective output transformer.

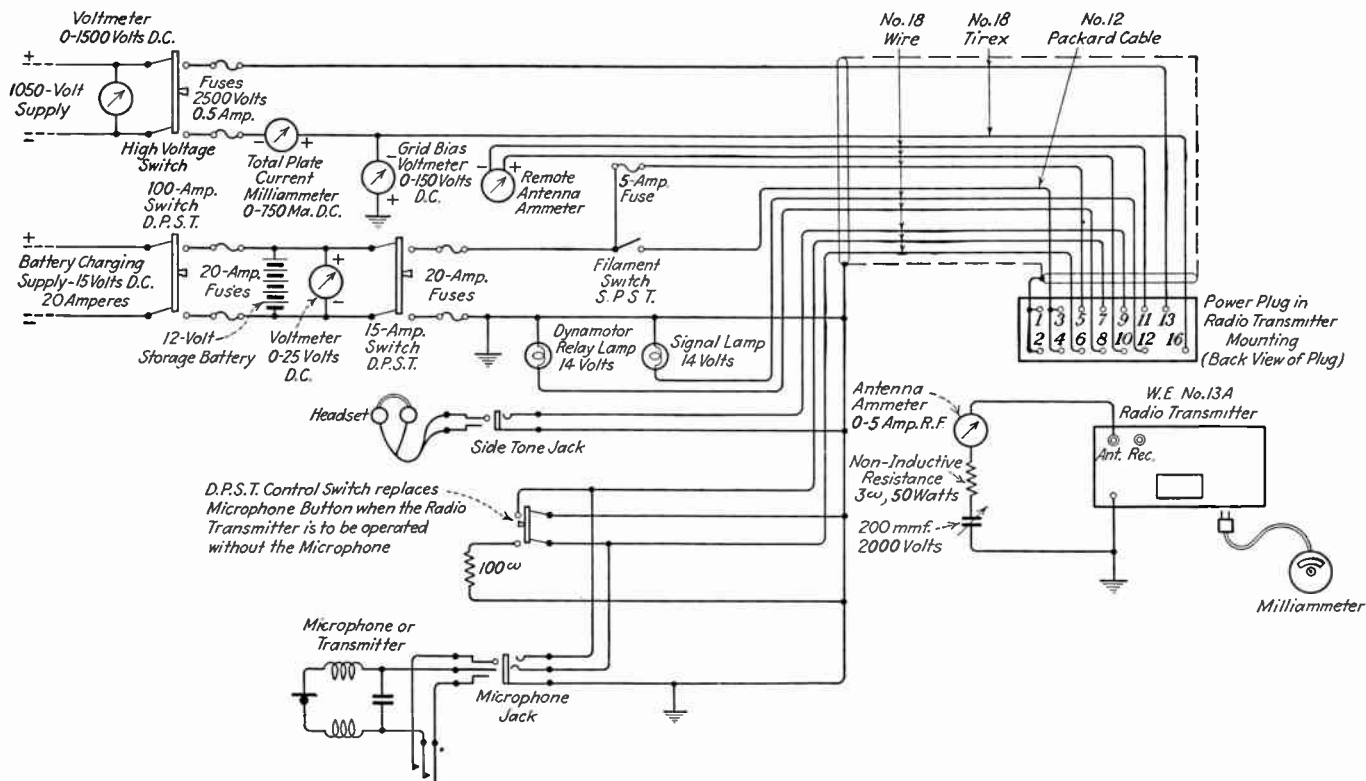


FIG. 278.—Test bench for aircraft radio transmitters. (Courtesy of Western Electric Company.)

If the radio-frequency output is abnormally high and the modulation poor, check the adjustment of the antenna coil. Measure the screen voltage. If it is high, the cause is probably that one or more of the radio-frequency amplifier tubes are drawing a very low or reversed screen current. Replace them one at a time until the one causing the trouble is located. If it is a second-amplifier tube it may give additional satisfactory service in the first amplifier.

Abnormal heating of the vacuum tubes might be due to a ground occurring at any point in the system on the negative high-voltage lead causing the bias to be removed from all the tubes which would allow them to draw abnormally large plate currents.

If only one of the second amplifier tubes is abnormally hot, the trouble is due to the two parallel tubes refusing to divide the load. Check the wiring to be sure both tubes are connected. Interchange tubes until two are found which will divide the load. The removed tube may be used in the first amplifier.

266. Aircraft Radio-telegraph Transmitter. *a. Development.*—The Pan American Airways¹ which conducts most of its flying over tropical waters found radio telephony unsatisfactory and uses radio telegraphy exclusively for all its radio-communication work. Although an operator must be carried, there are advantages to this method in that not so much shielding of the ignition system is required, accuracy and speed of message handling are increased, and ground stations do not need to be so close together. Communication is secured up to 1,800 miles.

To meet international regulations, aircraft radio equipments for use in overwater flights must be capable of working on 500 kc./s. as well as on the regular international air service calling frequency of 333 kc./s. In addition to these required intermediate frequencies, it was found that increased efficiency in the face of terrific tropical static could be secured on the high frequencies. Accordingly the transmitters used in this service are also designed to work on 3,090, 5,550, and 9,370 kc./s.

b. The Transmitter.—A circuit diagram of a radio-telegraph transmitter, very similar to the type used on aircraft of the Pan American Airways, is shown in Fig. 279. Because of the constant evolution and improvements in construction this circuit cannot be considered as actual in every detail. It is, however, representative of the transmitter used.

The transmitter is rated at 12 watts and utilizes two -10-type tubes in a master-oscillator power-amplifier circuit. The oscillator is of the split inductance or Hartley type. A distinctive feature of this transmitter is that plug-in inductance and capacity units are used for each frequency employed. An adjustable tank condenser is a part of each

¹ This system utilizes over 125 ground stations, has over 150 air liners equipped for radio telegraphy, and employs over 20 radio direction-finder stations.

plug-in unit, there being two units for each frequency, one unit for the oscillator and one for the amplifier, as shown in the diagram. The capacity required is achieved by proper adjustment of the unit when the set is tuned. After this initial adjustment it is unnecessary for the operator to tune the tank circuit when shifting frequency. The only tuning adjustment is the antenna oscillatory circuit (L_3C_3) which consists

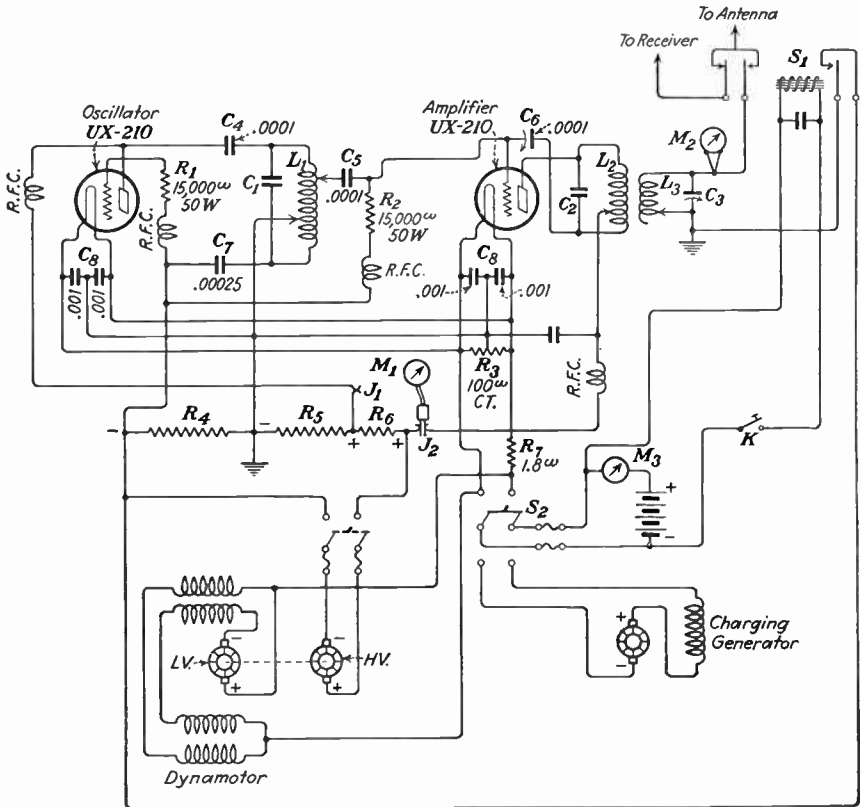


Fig. 279.—Typical radio-telegraph transmitter for aircraft.

of a tapped inductance and a variable condenser. Thus to shift from one frequency to another the operator simply inserts the proper plug-in units and resonates the antenna for proper radiation.

On the Pan American aircraft transmitters, used in the tropics, short- and medium-distance traffic is handled on a day frequency of 5,692.5 kc./s. and a night frequency of 3,082.5 kc./s. In addition to these frequencies, long-haul day traffic is handled on 8,220 kc./s. for distances of 500 or 600 miles, depending on conditions. Frequencies of 500 and

333 kc./s. are also provided as required by international law for aircraft flying over water.

Keying for radio telegraphy is accomplished by keying the grids of both the oscillator and the amplifier. It will be noted that when the key is up a blocking bias is developed across R_4 which stops operation of both tubes. When the key is down (closed), the grids are connected to the filament center-tap through their respective grid leaks and radio-frequency chokes as shown in the circuit diagram.

Plate current for both tubes can be read by the milliammeter M_1 which is fitted with a plug for insertion into either the oscillator or amplifier plate-circuit jacks, J_1 , J_2 . Antenna current is read from the thermocouple meter, M_2 . A zero-center ammeter M_3 is inserted in the positive storage-battery lead which reads "charge" or "discharge" current.

The relay S_1 provides break-in operation. When the hand key is pressed, all relay armatures are drawn toward the relay core. This action removes the blocking bias from the grids of both tubes by connecting the remote end of R_4 to the ground thus short-circuiting it; the inductance L_3 is connected to the antenna and the receiver is disconnected from the antenna. When the key is released, the relay armatures spring back disconnecting the transmitter from the antenna while connecting the receiver to it, and the blocking bias taken from across R_4 is applied to the grids of the tubes.

The primary source of power for the transmitter is a 12-volt storage battery which operates the tube filaments and the relay S_1 , and drives a dynamotor which delivers high-voltage plate potential. The rating of the dynamotor is 400/12 volts, 70 ma./8 amp. No filtering is required.

The storage battery is recharged by a small generator rated at 6,500 r.p.m., 14 volts, 5 amp. This machine is driven by either the engine of the plane or a wind drive.

The voltage-divider resistance is made up of units R_4 , R_5 , and R_6 . Unit R_4 provides blocking bias for both tubes for keying purposes, R_5 is a bleeder resistance, and R_6 drops the voltage to the proper value for the oscillator.

The transmitter can be started and stopped by the main power switch S_2 which also serves to put the battery on charge. The high-potential plate supply is controlled by a double-pole single-throw knife switch, and all lines are properly fused.

267. Ultra-short-wave Transmitters. *a. General Considerations.*—The last portion of the radio-frequency spectrum to be developed for communication purposes is that containing the ultra-short- or ultra-high-frequency waves below 10 meters; above 30,000 kc./s.

One of the first of these systems put into practical commercial use was installed by The Mutual Telephone Company for the Interisland Telephone System in Hawaii in 1931. Since then similar systems have been installed for general communication purposes by the U. S. Army and Navy, the state forestry services, the broadcasting companies for remote pick-up purposes, and a constantly increasing number of police depart-

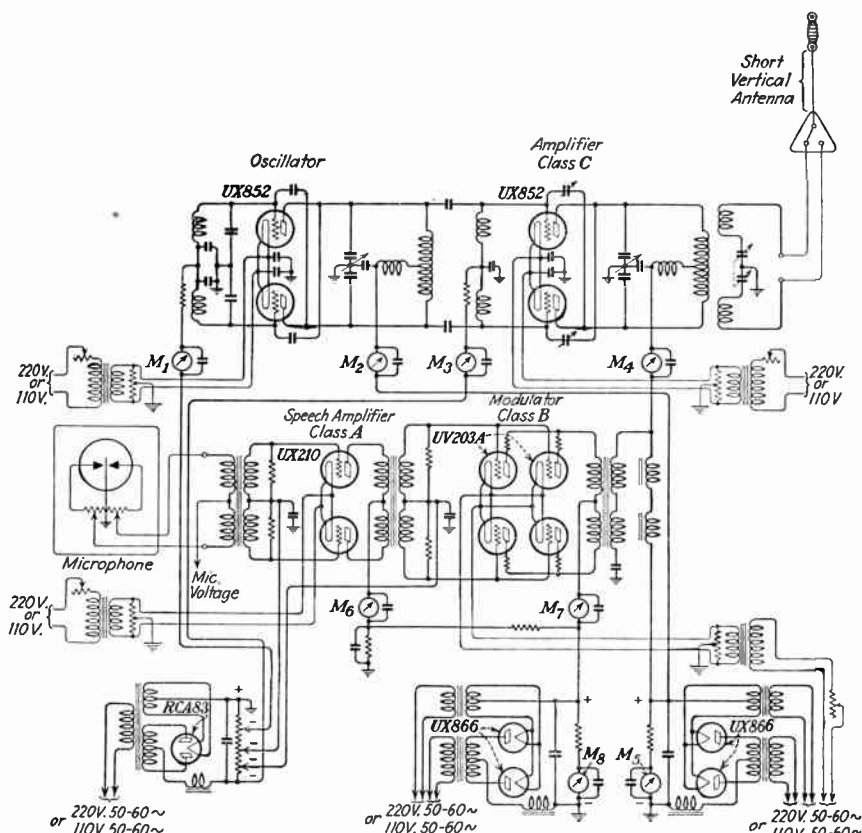


FIG. 280.—Ultra-high-frequency radio-telephone transmitter. (Courtesy of RCA Manufacturing Company, Inc.)

ments. The possibilities for further adoptions seem limitless owing to the restricted nature of the radiations employed.

b. A 100-watt Radio-telephone Transmitter: RCA Victor Type ET-5012.—This transmitter is designed to cover the frequency band of approximately 30 to 36 megacycles, but it is possible to operate the transmitter in either a lower or higher frequency band than that for which it was primarily designed by changing various inductance coils and the

antenna tuning to be in accordance with the desired frequency. A simplified schematic diagram of this transmitter is shown in Fig. 280; the control circuits have been omitted for simplification. A close study of this diagram will make clear the significant features of the circuit.

The power amplifier is operated as class C giving a high plate efficiency. Both the oscillator and the amplifier are self-biased, thus reducing the number of necessary adjustments. Neutralization is accomplished by means of an adjustment within the protective screening. Once this adjustment has been made, the neutralizing capacitor may be locked in place.

Three vernier dials are provided on the front panel which provide for all tuning necessary. These dials tune the oscillator, amplifier, and antenna circuits as shown in the diagram and are fitted with a locking device to insure that the transmitter will not accidentally be thrown out of adjustment.

Due to the use of the push-pull master oscillator, the frequency stability is well within the limits required for reliable communication in the ultra-short-wave spectrum.

Provision is made for 100 per cent modulation in this transmitter, thereby increasing effectively the range and the signal-to-noise ratio of the equipment. Accordingly the final power stage is specifically designed to handle 400 watts peak power without introducing distortion.

The audio equipment is designed to give faithful reproduction of all speech frequencies; that is, the frequency response is flat to within 2 db from 100 to 5,000 cycles. The audio-input circuit is designed to operate from a 500-ohm line at a level of +6 db (0 db = 12.5 mw. in 500 ohms), in order to obtain proper 100 per cent modulation. This transmitter operates with either the carbon-, condenser-, or velocity (ribbon)-type microphone.

All component parts of the radio-frequency circuit are insulated from their base by low-loss Isolantite supports. Isolantite antenna posts project through the top cover of the housing. All items, including the separate power-supply systems, are completely enclosed in shields to prevent pick-up.

The transmitter requires approximately 1.6 kw. from the line. During periods of standby between dispatches, the power consumption is reduced to approximately 400 watts.

The following meters indicated on the diagram are provided:

- | | |
|---|---|
| M_1 Grid current of master oscillator. | amplifier. |
| M_2 Plate current of master oscillator. | M_6 Plate current on speech amplifier. |
| M_3 Grid current of power amplifier. | M_7 Plate current of modulators. |
| M_4 Plate current of power amplifier. | M_8 Plate voltage on speech amplifier and |
| M_5 Plate voltage on oscillator and | modulator. |

The transmitter is usually placed in a room as near the antenna as possible, with access through a window for a pair of wires running to the antenna. The antenna is a single vertical wire, approximately 10 to 14 ft. long, drawn alongside a 25- to 40-ft. pole placed on the roof, or as high as possible.

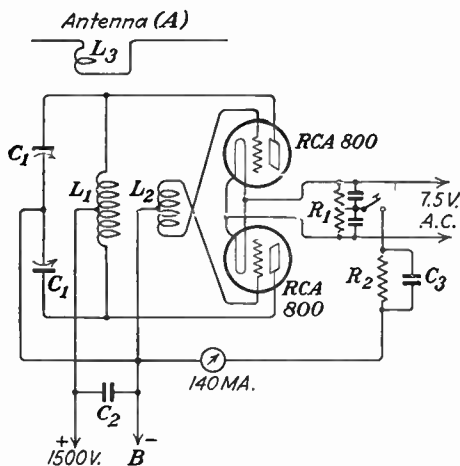


FIG. 281.—Five-meter push-pull oscillator circuit. (Courtesy of RCA Manufacturing Company, Inc.)

Because of the difficulty of frequency stabilization on the ultra-short waves, ICW emission is permitted. This is accomplished by using an alternating-current plate potential or by modulating the oscillator plate current by means of an audio oscillator as was described earlier in this chapter in connection with the intermediate-frequency marine transmitter.

Radio-telegraph transmitters are infrequently used on the ultra-high-frequency bands as the advantages of radio-telephony in the absence of static and over limited ranges more than outweigh the expense and difficulty of employing speech-modulated signals.

A very satisfactory 5-meter push-pull oscillator circuit, designed by the RCA Radiotron Division of the RCA Manufacturing Company Inc., and reproduced from their *Bulletin* 464 by permission, is shown in Fig. 281.

c. Radio-telegraph Transmitter.

Either the Gutton-Touly oscillator or the Mesney balanced oscillator, shown in Fig. 137, could be used for radio-telegraphic communication by coupling them to an antenna system. Keying may be accomplished by inserting a key in the grid return or by placing a radio-frequency choke in the grid circuit, shunting the key around the choke and thus accomplishing keying.

CHAPTER XI

RADIO RECEIVERS

In any consideration of receivers it must be kept in mind that the receiver constitutes one half of the radio-communication system and as such is designed to operate best from the particular type of transmitter which constitutes the other half of the system. This means that receivers must be designed with the requirements of the service in mind, and that, while some receivers are in some respects universal, most receivers are intended for a particular function, that is, to operate in a particular class of radio-communication service.

Radio receivers are classified in accordance with the conventional nomenclature applied to the radio spectrum as shown in Fig. 106.

I. GENERAL-PURPOSE RECEIVER

268. All-wave Receiver. *a. Description.*—The RCA Victor model AVR-5-A all-wave receiver (Fig. 282) has been designed to cover a frequency range of from 150 to 18,000 kc./s. in five bands as follows.

	Switch position (S_{2-7})	Wave band	Frequency range, kilocycles	Wave length, meters	
Airways weather	1	<i>X</i>	150- 410	2,000 -732	
Entertainment	2	<i>A</i>	540- 1,500	555 -200	
Aviation communication,	} 3	<i>B</i>	1,500- 3,900	200 - 77	
police reports, broadcasting,		} 4	<i>C</i>	3,900-10,000	77 - 30
amateurs, etc.			} 5	<i>D</i>	8,000-18,000

Any one of the frequency bands may be obtained by the turn of a single knob or station selector to the proper alphabetical selection of channels 1, 2, 3, 4, and 5. Scales 1, 2, and 3 are shown on an illuminated dial and are calibrated in kilocycles, the setting being indicated by the lower end of a moving pointer. The upper end of this pointer moves over the scales 4 and 5 which are calibrated in megacycles (1 megacycle = 1,000 kilocycles). The police bands, indicated by the letter P, and associated numerals followed by the letter M, represent the wave length in meters at certain points on the scale.

The receiver, as shown in Fig. 283, uses the superheterodyne principle of double detection, automatic volume control, manual volume control,

diode detection, and class B (push-push) power amplification. All tubes employed in the high-frequency portion of the receiver are of the triple-grid and pentagrid types.

The first tube (1) or radio-frequency amplifier is a triple-grid super-control amplifier of the -58 type. This tube has an amplification factor of 1,280 and serves to build up the incoming radio-frequency signal voltages to a very high degree. The second tube (2) is another -58

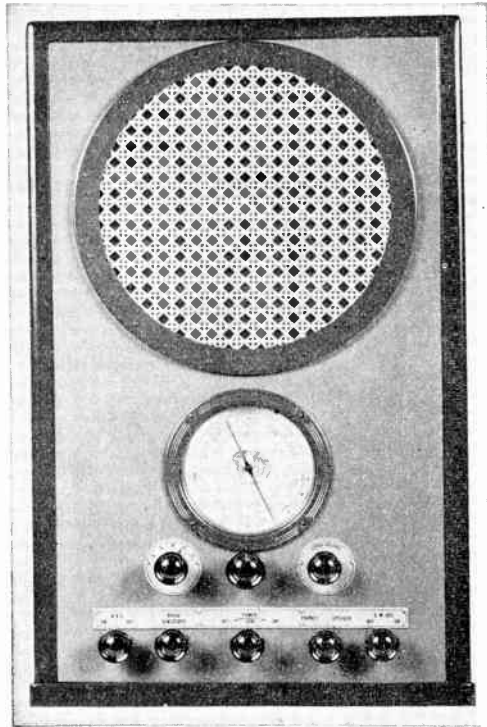


FIG. 282.—All-wave receiver, RCA Victor model AVR-5-A. (Courtesy of RCA Manufacturing Company, Inc.)

radio-frequency amplifier to increase still further the radio-frequency gain or signal voltages. The third tube (3) is a pentagrid converter of the 2A7 type and serves as a combination first detector and oscillator. The fourth tube (4) is another -58 radio-frequency amplifier which functions as an intermediate amplifier of the heterodyne frequency of 445 kc./s. The fifth tube (5) is a duplex-diode 2B7 type which serves several functions in one, namely, to rectify the 445-kc./s. signal frequency, to amplify the rectified variations, and to apply an automatic volume-control action to the radio-frequency amplifier tubes. The sixth tube (6) is a simple triode audio-frequency amplifier of the -56 type and serves to increase the audio-frequency voltage variation to the proper amplitude

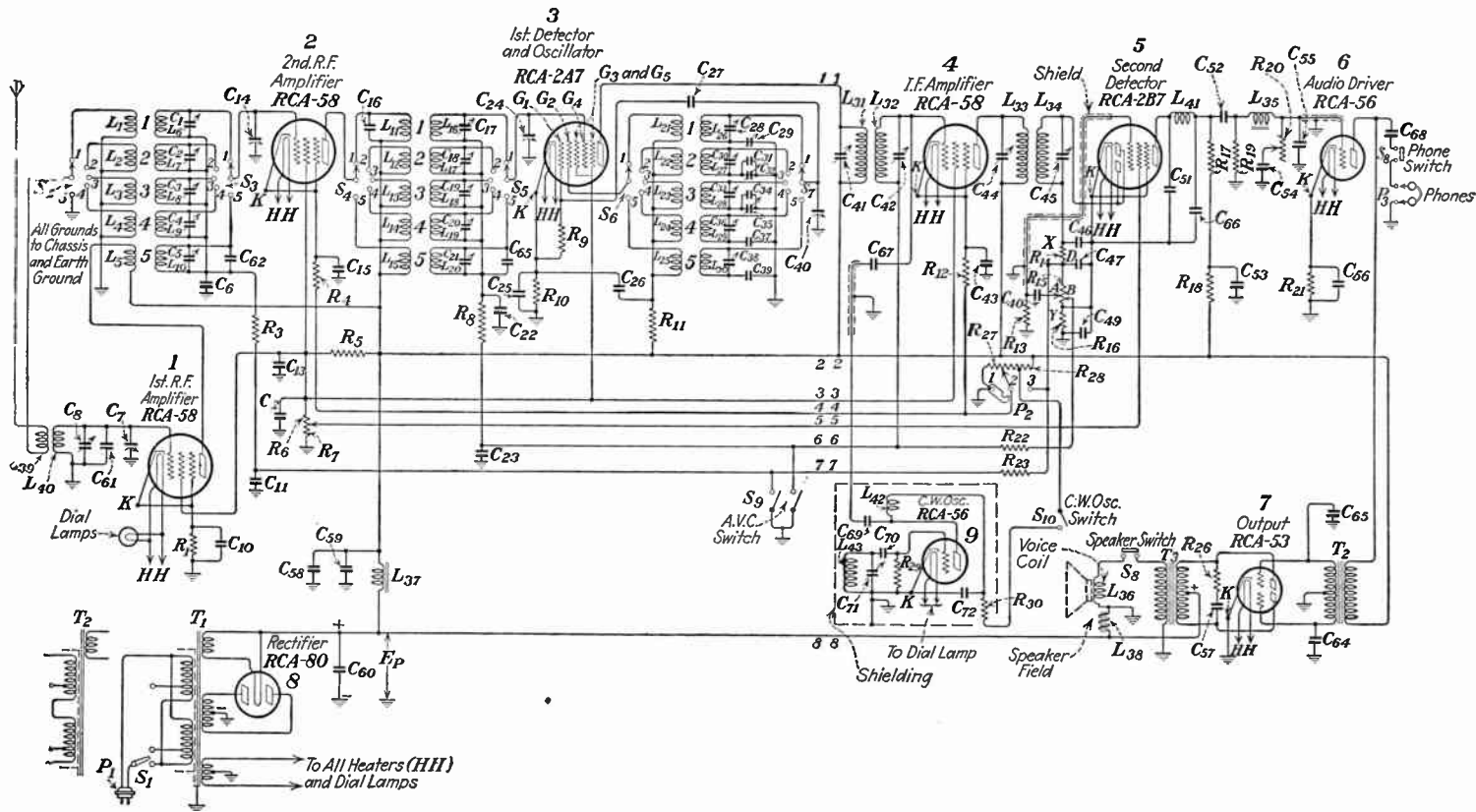


FIG. 283.—Circuit diagram, RCA Victor all-wave receiver, model AVR-5-A. (Courtesy of RCA Manufacturing Company, Inc.)

to drive the power amplifier. The seventh tube (7) is a twin power-amplifier -53-type tube which converts the audio-frequency voltage variations into power variations to operate the dynamic loud-speaker. This tube operates as a class B amplifier.

The eighth tube (8) is the full-wave rectifier -80 type which changes the stepped-up alternating-current supply voltage to direct current for the proper operation of all plate and bias circuits.

The ninth tube (9) is a second-triode oscillator -56 type which beats against the 2B7 second-detector tube for the reception of *CW* telegraph signals. The theoretical operation of this principle is explained in Sec. 157.

The beating effect in this case takes place in the plate circuit of tube (4), since the oscillator output is fed to its grid through the coupling condensers C_{67} and C_{69} . The combined *CW* intermediate and oscillator frequencies now developed across the detector input circuit will produce an audio frequency in the vicinity of 1,000 cycles to the audio circuit.

b. Theoretical Operation.—When the alternating-current line switch S_1 is closed, the primary winding of the power transformer is energized. A high voltage of an alternating character is developed across the two plates of the type -80 full-wave rectifier tube (8) having an r.m.s. value of 640 volts. Two low alternating voltages of 5 and 2.6 volts, respectively, are also simultaneously developed across the two small secondary windings which serve to heat the filaments of the type -80 rectifier and the amplifier tubes. The rectifier tube draws 2 amp. from the 5-volt winding, and the seven amplifier tubes and dial lights draw a little less than 8 amp. from the 2.6-volt winding.

The alternating-current voltage developed across the high-voltage winding is then rectified through the action of the full-wave tube in accordance with the general principles of rectification. Hence the filament of the -80 tube becomes the positive high-voltage lead which supplies direct current at the proper potential to the various tube plates and screens, the negative lead being the center-tapped or grounded section of the high-voltage winding. The direct-current supply is then suitably filtered to provide a smooth or unvarying plate and screen-grid current flow in all tube circuits by the filter system C_{59} , C_{60} , and L_{37} . The field magnet L_{38} of the dynamic speaker is energized by direct connection across the rectified high-voltage supply. The normal or static plate- and screen-current drain for all amplifier tubes is approximately 67 ma. during the period in which no signals are being received. This value may be increased to approximately 100 ma. when signals are being received owing to the characteristics of the type -53 power-amplifier tube which is adjusted as a class B amplifier.

The grid-bias voltages for all tubes are applied through the voltage drops developed across the resistances in series with the cathodes. The grid-bias voltages for the first and second radio-frequency amplifier tubes is produced by the resistances R_1 and R_4 , respectively; for the first detector-oscillator tube 3 through resistance R_{10} ; for the intermediate-frequency amplifier tube (4) through resistance R_{12} ; for the audio-frequency grid (control grid used here) of the second detector tube (5) through resistance R_{16} ; for the audio-frequency driver tube (6) through resistance R_{21} . No grid bias is required for the power-amplifier tube (7) when operating as a class B amplifier owing to its design characteristics. All condensers connected across these resistances are radio-frequency by-passes to provide paths of low reactance for the signal frequencies.

Resistances R_{16} and R_{21} have large values of capacitance connected across them to insure effective by-passing of the lower audio frequencies.

Grid- and plate-filtering resistances and condensers are provided in all important circuits in which the maximum circuit efficiency is required in accordance with the theory of grid- and plate-filtering principles.

With all the proper plate, screen, bias, and filament voltages applied, and plate currents flowing in all circuits, the operation of the receiver with an incoming signal is as follows.

A signal voltage induced in the antenna develops an alternating e.m.f. of high frequency in either coil L_1 , L_2 , L_3 , or L_4 and the series coil L_{39} , depending upon the position of the switch S_2 and the frequency of the signal being received. Let us assume that all switches S_2 , S_3 , S_4 , S_5 , S_6 , and S_7 are in position 1 which utilizes the band of frequencies from 150 to 410 kc./s. The only radio-frequency transformers in use when the switches are in this position are L_1L_6 , $L_{11}L_{16}$, and $L_{21}L_{26}$, their associated tuning condensers, and the main or dial tuning condensers C_{14} , C_{24} , and C_{40} .

Assuming a signal in the 150 to 410 kc./s. band is inducing an e.m.f. into the antenna circuit, a current flow of an alternating character of high frequency will flow through the primary windings L_{39} and L_1 of the two radio-frequency transformers. The resultant varying magnetic fields developed across these windings will induce an e.m.f. across the secondary windings L_6 and L_{40} provided that the LC constants of these two secondaries satisfy the conditions of resonance.

It will be noted from the switching arrangement that the plate circuit of the first radio-frequency tube (1) functions only into the primary winding of the coil L_6 which, together with its secondary L_{10} , is used for the high-frequency short-wave spectrum 5 of from 8,000 to 18,000 kilocycles. Consequently, with all switches in position 1, an incoming frequency in the 150- to 410-kilocycle spectrum will not cause the first radio-frequency tube (1) to function since the secondary L_{40} and the output circuits L_5 , L_{10} are tuned to the short-wave spectrum. Hence, the varying magnetic fields around L_{39} and L_1 will induce an e.m.f. *only* across the L_6 tuned circuit since it alone may be adjusted to resonance in this frequency band.

The radio-frequency voltages now developed across L_6 are applied directly across the GK terminals of the second radio-frequency amplifier tube (2) through the switch connection S_3 on contact 1 and the two by-pass condensers C_6 and C_{15} . The signal-voltage variations across GK produce resultant plate variations through L_{11} and return to the cathode through the radio-frequency by-pass condensers C_{58} and C_{15} . These increased magnetic variations now present across L_{11} are induced into the tuned circuit L_{16} and its associated tuning condenser which develops an e.m.f. that is applied across the control grid G_4 and the cathode K of the first detector and oscillator tube (3).

The input voltages to tube 3 produce resultant variations in its plate circuit and the primary winding of the intermediate-frequency transformer L_{31} which variations return to the cathode through the by-pass condenser C_{58} . These frequencies do not develop an e.m.f. across the secondary winding of this transformer since it is resonant only to the intermediate frequency of 445 kilocycles. Hence, the incoming frequencies do not flow effectively in the plate circuit owing to improper resonant conditions. It will be noted, however, that one end of the primary winding of the radio-frequency transformer L_{21} is connected to G_2 , while the other end of the winding connects to the positive side of the high-potential. This, in effect, makes the grid G_2 equivalent to a plate or anode since it is positive with respect to the cathode. Now the secondary winding of this same transformer, L_{26} , is connected through contact 1 through a grid condenser C_{27} to the grid G_1 of this tube. That is to say, the plate and grid coils L_{21} and L_{26} form the inductive feed-back system of the oscillator portion of the circuit by using G_2 as the plate of the oscillator and G_1 as the control grid of the oscillator. Oscillations are produced in accordance with the general principles of

vacuum-tube oscillators, the frequency of which is dependent upon the circuit constants L_{26} , C_{28} .

The local oscillations developed in this tube combining with the incoming-signal frequency produce a resultant frequency of an intermediate frequency of 445 kc./s. in accordance with the general theory of the heterodyne action in superheterodyne receivers. This intermediate frequency produces voltages and corresponding current variations in the plate circuit, causing magnetic variations to be developed around the primary winding of the intermediate-frequency transformer L_{31} . These 445-kc./s. variations develop an e.m.f. across the secondary winding $L_{32}C_{42}$ since this combination is resonant to that frequency. This e.m.f. is then applied across GK of the intermediate-frequency amplifier tube 4 which produces resultant variations in its plate circuit and through the primary winding of the second intermediate transformer L_{33} . This varying field develops an e.m.f. across the secondary winding $L_{34}C_{45}$. At this point the 445-kc./s. signal-voltage variations are rectified to audibility through the medium of the detector action of the tube. This tube (5) is of the duplex-diode-pentode type which serves the triple function of a diode detector, automatic volume control, and audio-frequency amplifier, all in one.

The detecting action takes place by connecting the two diode plates together and then connecting them to one end of the secondary winding of L_{34} . The other end of the secondary is connected through a condenser C_{46} to the cathode.

The signals developed across L_{34} are, therefore, applied directly across the diode plates and cathode of the tube. This results in asymmetrical pulsations or rectified variations to pass through the resistances R_{14} and R_{15} to the cathode in conformity with the theory of diode detection. In other words, these pulsations are now represented by the audio-frequency or rectified-signal variations now present across the resistances between the points X and Y . Hence, audio frequencies will be present owing to this detecting action, between the sliding arm A of the potentiometer and the cathode K of the tube or point Y . These audio-frequency variations are then applied to the control grid of this tube through the coupling condenser C_{40} which then functions as an audio-frequency amplifier tube. The grid leak R_{13} prevents the grid from blocking and also allows the grid bias developed by R_{16} to be applied to the audio-frequency grid which is the grid nearest the cathode. These audio-frequency variations are then present in the plate circuit and the audio-frequency coupling resistance R_{17} .

The intensity or volume, of these variations is determined by the position of the manual volume-control moving arm A with respect to the point Y or cathode. The closer A is to Y , the less will be the potential difference between the audio-frequency grid and the cathode, and the lower will be the audio-frequency voltages developed. It will be noted, however, that the amplitude of the audio-frequency voltages will also be governed by the AVC or throttling action upon the radio-frequency grids in accordance with the theory of the automatic volume control. That is to say, during the period in which audio-frequency variations are present in the resistances R_{14} and R_{15} , the point X will be negative with respect to the point Y (see Diode Detector, Index). Hence, the two points D and B are also negative with respect to Y . Thus the negative voltages developed at D , owing to the potential difference between the points D and Y when the rectified signal is flowing through the resistances, supply a supplementary negative bias to the grid of the second radio-frequency amplifier tube (2) through the time-delay and filter resistances R_{23} and R_3 , respectively. The contact B also supplies a supplementary or throttling bias to the grids of the first-detector and the intermediate-frequency amplifier tubes (3) and (4) through the time-delay and grid-filtering resistances R_{22} and R_3 , respectively.

The audio-frequency voltages then developed across R_{17} are fed across GK of the audio-frequency driver tube (6) through the coupling condenser C_{62} and the choke L_{35} .

A series-resonant circuit L_{35} , C_{54} , and R_{20} is introduced across GK of this tube to attenuate the higher audio-frequencies above 3,000 cycles to provide the desired tone-control action. The resistance R_{20} determines how much of the higher audio frequencies are to be attenuated.

The audio-frequency variations now present in the plate circuit and the primary winding of the audio-frequency transformer T_2 are then induced across the secondary winding of this transformer and to the two grids of the twin-amplifier tube (7). The audio-frequency voltages developed here apply alternate charges to the two grids, namely, when one grid is positive (+), the opposite grid is negative (-).

This tube, unlike the ordinary class A amplifier, does not require a negative bias when the amplifier is, as in this receiver, operated as class B. When no signals are being applied, the $I_p E_c$ characteristic curve is such that at zero grid bias very little plate current flows, so that in effect the result is the same as operating near the cut-off point, as with the usual biased class B amplifier. The plate current under these conditions is approximately 18 ma. per plate, or 36 ma. for the two plates, passing out of the center-tapped transformer winding.

When a signal e.m.f. is applied across the secondary of T_2 the grids of both tubes draw grid current, and the plate current rises to a definite amount depending upon the value of the input-signal voltage. For example, when an audio-frequency signal of 8 volts (r.m.s.) is applied to the grids, each tube draws a grid current of approximately 3 ma. and a plate current of approximately 25 ma. At this value the output power developed across the load impedance T_3 is approximately 10 watts with a load impedance of 8,000 ohms measured from plate to plate.

The input transformer T_2 is of special design to fulfill the conditions required in an unbiased class B amplifier, namely, to provide the necessary excitation voltage and the power which is consumed by the grids when current flows. Hence a step-down transformer of an approximate ratio between 2:1 and 5:1 is used.

The audio-frequency variations then developed in the primary winding of the output transformer T_3 induce an e.m.f. across its secondary winding which energizes the voice-coil winding of the dynamic speaker. This winding is of low impedance to match the voice-coil impedance, so that a maximum degree of power may be absorbed by the load in accordance with the laws of impedance matching. The audio-frequency signal variations now present around the voice coil L_{36} act and react upon the steady field of the dynamic speaker "pot" magnet L_{38} , causing it to vibrate and produce audible sound variations through the medium of the attached cone.

The operation of the receiver is the same regardless of the position of the wave-changing switch, with the exception that, when the switches are in the 5 position, an additional stage of radio-frequency amplification (1) is provided to increase the sensitivity and selectivity on the high-frequency band, 8,000 to 18,000 kilocycles.

In this position (5) the incoming signal is induced from L_{39} to L_{40} and to the grid-cathode of the radio-frequency amplifier tube (1). These voltages then produce current variations in the plate circuit through L_5 which are then induced across L_{10} and applied to the grid-cathode of the second radio-frequency amplifier tube (2) and then amplified as previously explained. If it is desired to receive CW telegraph signals the oscillator switch S_{10} is closed and the speaker switch S_8 is opened. Reception may then be obtained by closing the phone switch S_8 and inserting a pair of high-resistance telephones in the output circuit of the -56 driver tube. Pin jacks are provided for this purpose. Increased sensitivity or radio-frequency gain may be obtained by removing the AVC throttling action by closing switch S_9 and adjusting the bias on the second radio-frequency tube (2) by means of the control potentiometer P_2 . This enables the proper adjustment for either an increased radio-frequency gain or the suppression of inter-carrier noise when the AVC is not in use.

c. *Vacuum Tubes Employed.*¹—Most of the tubes used in this receiver are of the new pentode and pentagrid variety. The two radio-frequency stages employ a -58 type or triple-grid supercontrol amplifier tube. The theoretical action of this tube is similar to that of the usual screen-grid tube with the exception that it develops a very high amplification factor, or μ , almost double that of the ordinary screen-grid tube. This is accomplished by special design and is particularly due to the *suppressor grid* located next to the plate. This grid reduces the secondary emission from the plate and greatly increases the effective electron flow in the plate circuit, thereby raising the efficiency of the tube as a voltage-amplifying device.

The first detector-oscillator tube type 2A7 is of the pentagrid variety, generally known as a *pentagrid converter*. This tube has four grids, a cathode, and a plate. The grid adjacent to the cathode is grid 1 and serves as the control grid of the oscillator portion of the circuit. Grid 2, which is adjacent to grid 1, is used as an anode grid to take the place of the plate of the oscillator portion of the circuit. Grids 3 and 5 are screen or shield grids which envelop the control grid 4. Grid 4 is the regular control grid connecting to the cap of the tube which supplies the incoming-signal voltage to the tube. The screen grids 3 and 5 with their positive potentials act as an electrostatic shield against the effect of the other grids, which serves to increase greatly the output impedance R_p , thereby enabling much greater output voltage to be developed across a properly matched load impedance. Incidentally, the portion of this grid nearest to the cathode also reduces the effect of reradiation by its effect of reducing the inter-electrode capacitance.

This tube is particularly adaptable to the *electron-coupled* principle of oscillation, since the oscillator grid 1 and the anode grid 2 are located directly between the cathode and the control grid 4. Hence, when the local oscillations are produced between the oscillator grid and the anode grid, these oscillations vary the amplitude of the normal electron flow between the cathode and the control grid, due to the incoming signal, and the resultant bombardment of the electrons on the plate will be the difference between the two frequencies, namely, the oscillator frequency and the incoming-signal frequency. Thus the resultant variations in the plate circuit will be the heterodyne frequency of 445 kc./s. which results in electron variations of this frequency in the plate circuit and L_{31} .

The second-detector tube type 2B7 is of the duplex-diode-pentode type which serves as both a diode rectifier and audio-frequency amplifier and also provides a screen and suppressor grid to increase its sensitivity as a voltage amplifier. The grid adjacent to the plate is the suppressor

¹ In studying the following descriptions, reference should be made to the tube characteristics and socket charts given in Appendix A.

grid which is internally connected to the cathode to put it at low potential with respect to the plate, thereby reducing the effect of secondary emission.

The grid adjacent to the suppressor grid is the conventional screen grid which serves to decrease the internal space charge. The grid nearest to the cathode is the control grid which connects to the cap of the tube and which serves as the audio-frequency amplifier grid. The two diode plates are connected together to form the rectifying portion of the tube.

The driver, or second, audio-frequency, amplifier tube type -56 is an ordinary triode-cathode-heater tube.

The output power tube of the -53 type is a twin amplifier using two grids and two plates which may be used as a class B push-push power amplifier or as a class A power amplifier. In the latter service the two grids and the two plates are tied together at the socket and operate equivalent to two tubes in parallel.

II. MARINE RADIO RECEIVERS

269. Marine receivers are especially designed for use aboard ship and incorporate features necessary for this service. They are primarily designed to receive radio-telegraphic signals and are very selective and sensitive. They cover the frequency bands allotted to marine work, and for this reason may also be used in shore stations which are the land terminals of the marine radio-communication systems.

Owing to the various conditions of climate and weather in which marine receivers are used, and to reduce receiver failures, it is necessary that they be ruggedly built. Simple circuits are favored so that repairs can be made, when necessary, by the operator at sea, away from service stations.

Long before the advent of broadcasting, marine receivers had reached a high state of development. The rapid perfection of broadcast receivers was due, in no small measure, to the experience gained by radio science in the building of efficient marine receivers. While improvements have been made in marine receivers since the beginning of broadcasting, these improvements are due mostly to the new types of vacuum tubes made available for broadcast receivers which have also lent themselves to the improvement of marine receivers. Certain special features, such as volume control and alternating-current operation, have been borrowed from broadcast-receiver design, but, on the whole, marine receivers constitute a class by themselves incorporating features which have grown out of the practical use of these receivers aboard ship.

As stated before, marine receivers are essentially designed for the reception of radio-telegraphic signals. It is true that radio-communication systems have been designed for the use of harbor tugs and other

small craft using radio telephony, but these installations have not yet reached important proportions, and, although classed as marine, they would perhaps be better classed as harbor radio systems to differentiate them from radio systems used on ocean-going and Great Lakes vessels. It is to this latter class that the classification *marine* is applied in this book.

As pointed out in the explanation of the regenerative and superheterodyne receivers in an earlier chapter (Secs. 155-157), it is necessary to produce beat frequencies to make the continuous-wave signals audible in the headphones. This is accomplished by using the regenerative circuit as an autodyne, or by using a superheterodyne receiver. There are other ways of making *CW* signals audible at the receiver, but these two methods are the only ones in practical use on marine frequencies.

As before explained (in Sec. 252 on transmitters) the emitted wave may be broken up into audio-frequency wave trains or interrupted continuous waves (*ICW*) and could thus be heard on any receiver whether the receiver was capable of producing a beat frequency or not, because the incoming signal would be modulated in character. This applies to all continuous waves regardless of how they are generated, be they from a vacuum-tube or an arc transmitter.

270. The Coupling Adjustment.—In marine receivers of the intermediate-frequency type which must be capable of receiving spark signals, which may be used aboard certain ships until 1940, it is the usual practice to provide inductive coupling between the primary and secondary circuits of the tuner. Owing to the broad frequency characteristics of spark emissions, it is necessary to use a minimum or loose coupling after communication has been established. On the other hand, it is the practice when using this type of receiver to use tight or maximum coupling for general listening-in purposes. This is known as the *standby circuit*, and this method of operation is followed at all stations, regardless of the type of waves being listened to, if an inductively coupled tuner is used. Any type of untuned antenna circuit which would make the receiver responsive over a wide band of frequencies could be referred to as a standby circuit. The establishment of definite calling and working frequencies by international agreement is making the standby circuit less necessary.

271. Audio-amplifier Response Characteristics.—Inasmuch as the marine radio-telegraphic receiver has to respond only to code characters sent or received on a very limited number of specific audio frequencies, and, in fact, that it is designed to be extremely selective in its tuned circuits, the audio transformers used with these receivers have a *peaked* frequency characteristic. They are known as *peaked transformers* and give highest relative amplification usually at some frequency between 500 and 1,000 cycles.

The characteristics of a peaked transformer which is peaked for 500 cycles is shown in curve *A*, Fig. 284. The frequency-response characteristic of an audio transformer designed for radio-telephone reception is shown as curve *B* of the same figure, so that the two types of transformers may be compared. It is readily seen that while the amplification of the transformer represented by curve *A* is not much different from the curve *B* transformer at frequencies around 100 and 2,000 cycles, the response of the peaked transformer at 500 cycles is much greater than that of the flat-characteristic transformer. Hence, a marine receiver equipped with ordinary broadcast-receiver types of audio transformer would not be

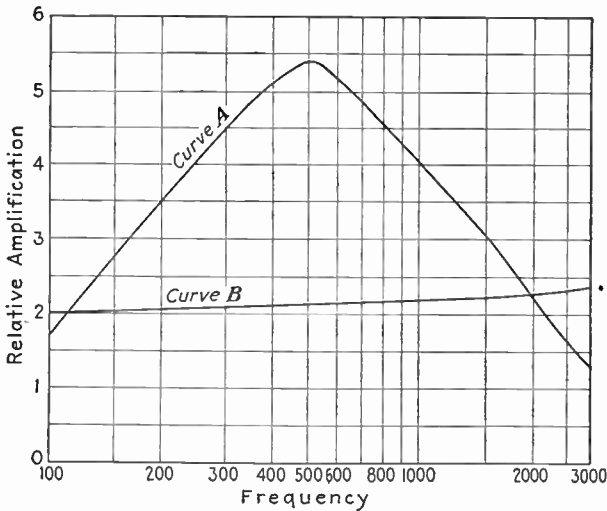


Fig. 284.—Characteristics of a peaked audio-frequency transformer.

very sensitive nor would it amplify the desired audio frequency and attenuate all other frequencies as the peaked transformer does.

272. Audio-amplifier Troubles.—If trouble in the receiver is localized to the audio-frequency amplifier and then further traced to a particular stage of audio amplification, it is quite likely to be due to some constructional or electrical defect in the audio transformer. Perhaps the most common cause of audio-transformer trouble in marine receivers is due to the effect of salt air on the apparatus. In spite of the best precautions to shield the receiver from climate, salt air is the surrounding medium and may cause trouble. The most common trouble from this source is corroded terminals on the transformer at the point where the soldered connection is made. Fiber insulation is generally used in audio transformers, and fiber has a certain amount of acid in it which, when aided by the salt air, may cause a weakening in the insulation of the transformer as a whole.

Another common trouble is that caused by surges due to the opening of the audio circuit in some manner, when full voltage is on. This might happen when shifting headphones from one amplifier stage to another, if automatic filament jacks are not provided to interrupt the current. Such surges cause arcing and may result in complete insulation breakdown and burning off of the fine wire used in the transformer winding; usually a No. 40 B. and S. gage wire is used.

The transformer core may sometimes be grounded causing an electrolysis to take place between the winding (anode) and the core (cathode), resulting finally in a burning out of the transformer winding affected.

Note: The primary and secondary circuits of an audio transformer may be tested by a pair of headphones and a 22½-volt B battery. The headphones and the battery are connected in series and then shunted across the winding to be tested. If the primary circuit is tested, no click will be heard if the winding is open (unless a by-pass condenser is connected across it). If the secondary winding is to be tested, no click will be heard if the winding is open. It is very important when testing the secondary to note carefully the intensity of the click, because the comparatively high resistance of the secondary winding will make the click much feebler than that heard through the primary. High-resistance phones should be used in these tests.

273. Shortening Condenser.—When it is desired to tune a receiver below the natural wave length of the antenna, a series condenser usually of the variable-air type is connected in the primary or antenna circuit. This condenser is variously known as a *short-wave* condenser (because its first use was to tune marine receivers down to the low waves when they were connected to large ship antennas), a *shortening* condenser, or as a *series-antenna* condenser. Regardless of its name its purpose is clear. In some receivers a switch is provided to short-circuit this condenser when it is not needed, in other receivers used with large antennas the shortening condenser is always in the circuit. If it is not desired to tune to signals below the natural wave length of the antenna provided, a shortening condenser is generally not included in the receiver circuit.

274. Power Supplies for Receivers.—Marine receivers are generally operated on batteries, using storage batteries for filament power and the conventional heavy-duty B batteries for plate power. Dry-cell type C batteries are also used.

The storage-battery equipment, including the special charging panel used, is described in Sec. 416 in the chapter on Storage Batteries.

Marine receivers operating on alternating current have been developed where the primary source of power is a small 110-volt alternator. A regular 280-tube rectifier is built in as an integral part of the receiver. Such an arrangement eliminates batteries entirely and simplifies the equipment somewhat.

275. Protective devices to protect the receiver against the damaging effects of near-by transmitters or heavy local atmospheric discharges are

provided in the form of a safety gap which is placed across the antenna and ground terminal posts on the receiver. This gap is set to arc over below the breakdown voltage of any insulation in the receiver. Therefore, any high-voltage surges or induced currents which might injure the receiver are automatically conducted directly to the ground. This gap, as pointed out, acts as a lightning arrestor when necessary.

When the transmitter is located close to the receiver, as on board ship, an automatic relay is provided which opens the connection from the receiver to the antenna when the transmitting key is pressed and makes the connection again when the key is released, constituting in effect a protective device and a break-in system which enables the operator to listen to signals when his key is up.

276. Multiple Reception.—When a number of receivers must be used on one antenna as is sometimes necessary, receivers utilizing one or more stages of radio-frequency amplification are used. Any local beat frequencies generated in a receiver are blocked by its radio-frequency stages, and all the receivers can be operated without the adjustments of one disturbing the adjustments of the others. Under these conditions a separate receiving antenna is generally used which can be much smaller than the main antenna owing to the high amplification of the received signal which is possible when radio-frequency amplifiers are used.

The manner in which the principles outlined thus far are applied to actual receiver design will become apparent from a study of the specific receivers which are described in the following sections.

277. High-frequency Receiver. *a. Description.*—The U.S. Coast Guard receiver, type CGR-26-1 is designed for use on vessels or at shore stations of the U. S. Coast Guard and has a frequency range of from 2,000 to 23,000 kc./s.

A set of plug-in coils is provided for the entire frequency range as follows:

- 2,000– 3,300 kc./s.
- 3,300– 5,500 kc./s.
- 5,500– 9,100 kc./s.
- 9,100–15,000 kc./s.
- 15,000–23,000 kc./s.

b. Antenna.—Because of the use of the coupling tube, the antenna requirements for this receiver are not very rigid. Almost any length of antenna will do, but for best operation an outdoor antenna of between 30 and 100 ft. in length is recommended. If the receiver is used in a locality where there is much interference, a short antenna should be used. Otherwise the long length will be found more effective.

c. Circuit.—The circuit arrangement of this receiver is shown in Fig. 285. The radio-frequency input to the receiver is led directly to the grid of the radio-frequency coupling tube (RCA-236) which has a resistor (R_1) to ground to furnish the necessary bias for the grid. This aperiodic input circuit allows the use of a variety of antennas.

The plate circuit of the coupler tube is tightly coupled to the tuned grid circuit of the detector so that as much of the detector grid-input impedance as possible is reflected into the coupling-tube plate circuit, thereby providing maximum gain with this type of circuit.

The detector (RCA-237) utilizes the grid-rectification principle. Regeneration is accomplished by inductive feed back controlled from the front panel by a potentiometer (R_5) which varies the detector plate voltage. A choke (L_6) in the detector plate lead prevents radio frequency from getting into the audio circuits.

The audio-frequency amplifier is transformer coupled (T_1 , T_2); output jacks are provided so that the headphones may be plugged into the plate circuit of either the first or second audio-frequency stages. An RCA-237 tube is used for the first-stage audio amplifier and a UX-112-A is used in the second stage.

The audio transformers (T_1 , T_2) are enclosed in a single case and are thoroughly impregnated to insure reliable operation under adverse temperature and humidity

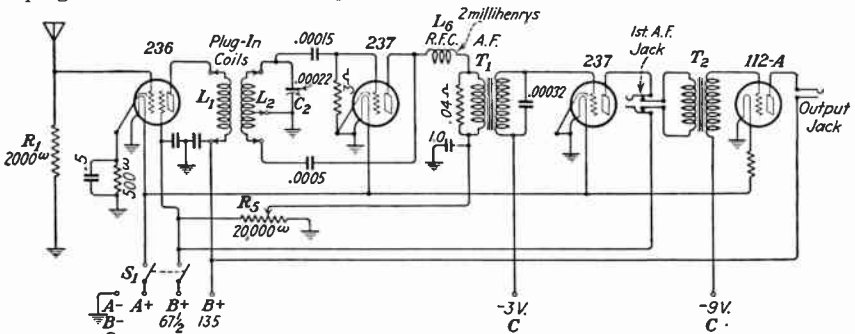


FIG. 285.—U. S. C. G. high-frequency receiver, type CGR-26-1. (Courtesy of U. S. Coast Guard.)

conditions. The transformers are designed to provide a substantially flat output characteristic between 150 and 3,500 cycles.

The supply voltages are as labeled in the diagram and are brought to a cable-connecting jack at the rear of the chassis. The various voltages are adequately by-passed wherever necessary.

To change from 6- to 32-volt filament supply or vice versa it is only necessary to remove the cover from the battery terminal block and change the link arrangement to correspond to the power supply available.

d. Operation.—The controls, located from left to right on the front panel, and their functions are as follows:

1. *On-Off Switch.*—In the OFF position, this double-pole, single-throw switch (S_1) opens both the filament supply, and the detector plate and radio-frequency screen-grid circuits.

2. *Tuning Control.*—This control (C_2) operates the single variable condenser that tunes the detector grid circuit.

3. *Audio-output Jacks.*—Just below and to the right of the tuning control are two phone jacks marked 1 and 2, corresponding to the first and second audio-frequency stage.

4. *Regeneration Control.*—The regeneration control (the small knob R_5 on the panel to the right of the tuning control) varies the voltage on the plate of the detector tube from 0 (at the extreme left) to plus $67\frac{1}{2}$ volts (at the extreme right), thereby controlling effectively the regeneration of the detector.

The controls are all arranged to increase their ultimate effect by motion clockwise, or from left to right. The tuning-control marking increases numerically with clock-

wise rotation of the knob and with an increase in frequency. The regeneration control increases the amount of feed back or regeneration with clockwise rotation of the knob.

To tune in a certain continuous-wave signal proceed as follows: Select the coil which covers the desired frequency range, turn the regeneration control (R_s) to the extreme left, turn the switch (S_1) to the ON position, insert a pair of high-resistance headphones in the audio-output jack 2, then with the right hand increase the regeneration control until a slight breathing or hissing sound is heard in the phones. With the left hand vary the tuning control slowly over the frequency band, at the same time moving the regeneration control with the right hand so that the hissing sound is just heard. This means that the detector is oscillating and is operating at its most sensitive point. After the beat note between the desired signal and the oscillating detector is heard, carefully adjust the regeneration control for maximum signal and move the tuning control until the desired pitch note is heard. A 1,000-cycle note will be found best because the headphones are most responsive to this frequency.

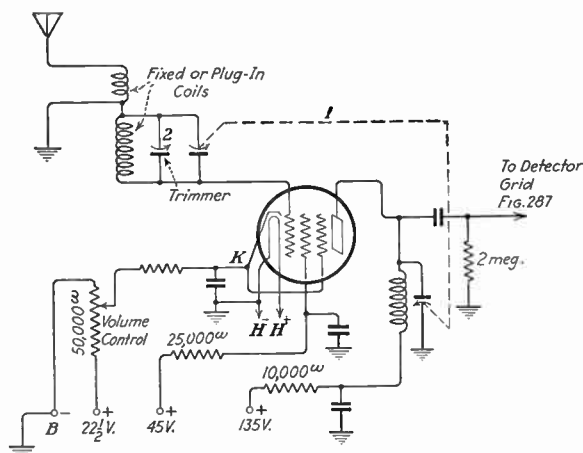


FIG. 286.—Radio-frequency section of R.M.C.A. model AR-8500 receiver.

To tune in a modulated or telephone signal, proceed with the above instructions except, after the signal has once been found, turn the regeneration control to the left until the detector is just below the oscillating point.

278. Combination Frequencies Receiver. *a. Description.*—The model AR-8500 receiver was designed primarily for marine applications and is supplied by the Radiomarine Corporation of America. Several important features not heretofore available are provided, which makes this receiver especially suited to modern marine traffic conditions. The manufacturer's description follows. A fixed-coil system covering the intermediate frequencies used in marine service (from 230 to 520 kc./s.) is provided, and in addition a series of plug-in coils, one set of which is used at a time and which can be immediately thrown into the circuit by a pull-pull button switch, is provided. A schematic diagram showing the way in which the tuning inductances are connected ahead of the first radio-frequency tube is shown in Fig. 286, while a complete wiring diagram appears in Fig. 287.

b. Frequency Range.—The receiver is designed to cover a total frequency range of 17,500 to 15 kc./s. (17.1 to 20,000 meters). For average marine service the following plug-in coils may be supplied.

Coil	Frequency, kilocycles	Wave length, meters
<i>A</i>	11,000–17,500	27.27– 17.14
<i>B</i>	8,200–12,900	36.59– 23.3
<i>C</i>	5,500– 8,900	54.55– 33.71
<i>H</i> (built-in fixed coils)	230– 520	1,304 – 577
<i>I</i>	110– 230	2,727 – 1,304
<i>J</i>	55– 110	5,455 – 2,727
<i>K</i>	30– 55	10,000 – 5,455
<i>L</i>	15– 30	20,000 –10,000
Special plug-in coils may be ordered to cover the following ranges		
<i>D</i>	3,500– 5,500	85.7 – 54.55
<i>E</i>	2,200– 3,500	136.4 – 85.7
<i>F</i>	1,100– 2,200	272.7 – 136.4
<i>G</i>	500– 1,100	600 – 272.7

c. Sensitivity.—The combination of a tuned radio-frequency amplifier, screen-grid detector, and pentode audio amplifiers permits a high degree of sensitivity to be obtained. In the high-frequency range of the receiver an input of 1 to 2 microvolts ($\mu\text{v.}$) will provide a good headphone signal. In the intermediate-frequency part of the range where modulated signals are frequently used, the receiver will provide a readable output with a 60-per cent modulated signal having a value of $2 \mu\text{v.}$ At the low-frequency portion of the range, satisfactory *CW* reception may be obtained with signals having a level of only 2 to $4 \mu\text{v.}$ Since the noise level, due to static, induction, etc., is usually greater than the input values mentioned above, it will be evident that the receiver provides ample sensitivity.

d. Selectivity.—The selectivity characteristics of the receiver may be determined by considering the following typical examples. Assume the receiver is adjusted to 500 kc./s. with critical regeneration to receive a $2\text{-}\mu\text{v.}$ modulated signal. An interfering signal 10 kc./s. lower than the desired signal must have a field strength approximately 50 times greater than the desired signal before it begins to produce barely audible interference. Under the same conditions as above but with the interfering signal 20 kc./s. lower, it must have a field strength 400 times greater than that of the desired signal.

At 125 kc./s. with the receiver adjusted for best regeneration and with a $1\text{-}\mu\text{h.}$ input, an interfering signal separated only 5 kc./s. must have a field strength 200 times greater than the desired signal before it produces audible interference.

At 8,250 kc./s. receiving a modulated signal, an interfering signal, 20 kc./s. “off” resonance, must be approximately 6 times stronger than the desired signal before it produces audible interference. At 40 kc./s. “off” resonance, the interfering signal must be approximately 20 times as strong as the desired signal to produce noticeable interference.

All of the above data are based on an adjustment where the detector is not oscillating. If an oscillating detector is used, such as is necessary for *CW* reception, the selectivity will in general be even better than that indicated above.

e. Receiver Circuits.—The complete circuits of the receiver are shown in Figs. 286 and 287. The input coil for the grid of the radio-frequency amplifier contains a

coupling coil and a tuned grid coil. The coupling coil is designed for each frequency band so that the coupling increases with wave length. The plate circuit of the radio-frequency amplifier is also tuned and is coupled through a condenser to the grid of the detector. A range-selector switch (labeled *Switch knob*) controls the various circuits for the plug-in coils and the fixed coils. A tickler coil is coupled to an amplifier plate coil and regeneration or detector oscillation is controlled by means of a variable condenser. Particular care has been directed in the design of this part of the circuit with the result that smooth regeneration is obtained, and the use of excessive regeneration will not cause undesirable howling in the headphones. The audio voltage built up in the detector plate circuit appears across a 50,000-ohm resistor and is then coupled through a low-pass filter network to the grid of the first audio amplifier.

Resistance coupling is used between the plate circuit of the first audio stage and the grid circuit of the second audio stage. The output from the -38 second-stage ampli-

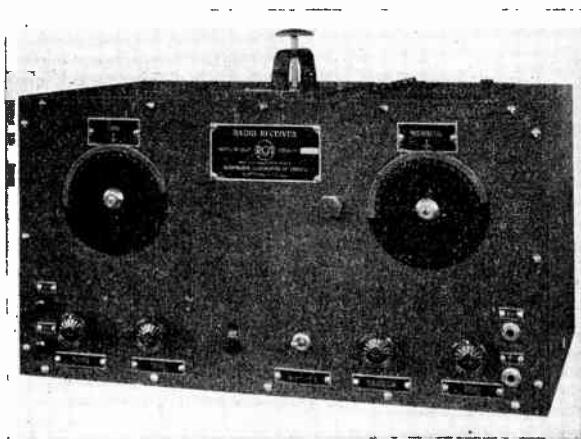


FIG. 288.—Combination-frequency receiver, R.M.C.A. model AR-8500. (Courtesy of R.M.C.A.)

fier is connected to the *high* telephone jack through an output transformer. The output of the first audio stage is connected to the *low* jack through a coupling condenser. No direct-current voltages exist on either phone jack.

f. Operation.—While the general operation of the receiver is quite similar to that usually employed in regenerative receivers, there are certain points in the model AR-8500 (Fig. 288) which deserve special attention.

Careful adjustment of the amplifier-trimmer condenser and the regeneration condenser will be found to produce excellent performance when receiving weak signals under conditions of interference. In any receiver used for *CW* work with an oscillating detector, it is necessary, of course, for the detector to be adjusted to a frequency differing by 500 or 1,000 cycles from that of the radio-frequency amplifier. In other words, the detector circuit is never tuned to exactly the frequency of the incoming signal, for, if this were done, zero-beat reception would result. Consequently in the model AR-8500 there has been provided a trimmer condenser for the amplifier grid so that this circuit may always be tuned exactly to the desired signal, while the detector is tuned slightly "off" to produce an audible beat note.

It will be found that the best setting for the regeneration condenser is influenced by the adjustment of the trimmer condenser. The rule to follow is always to use

sufficient regeneration so that the amplifier-trimmer condenser may be tuned through maximum without causing the detector to stop oscillating or drop into zero beat. When searching for a signal the correct adjustment of the regeneration control and the trimmer condenser may be checked easily by noting the increase in noise level as the trimmer condenser is turned through resonance.

On the front panel of the receiver, to the left of the regeneration knob, a small switch marked *G P* controls a resistor in the cathode circuit of the detector and provides in the *G* position grid-leak detection for small input signals and in the *P* position grid-bias detection for large input signals. In general, for maximum sensitivity this switch may be left in the *G* position. When operating on intermediate

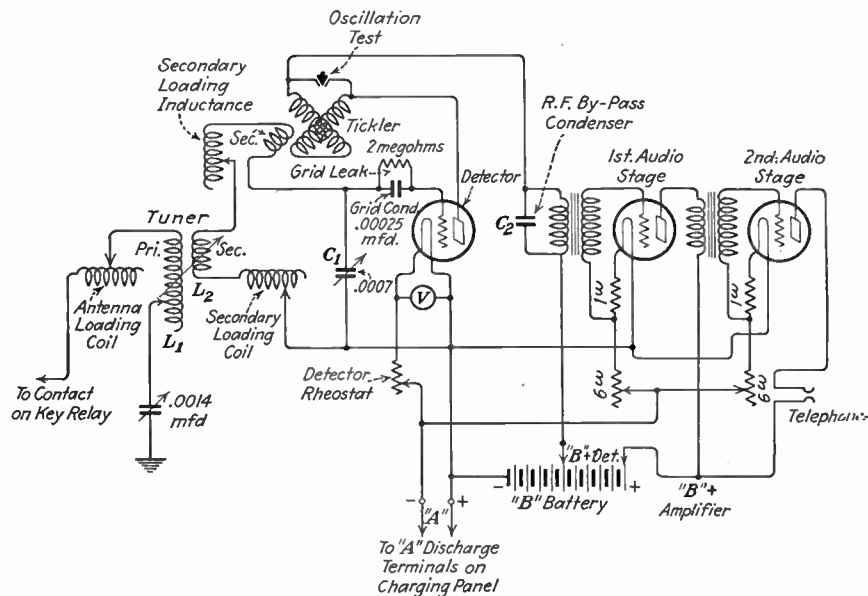


Fig. 289.—Schematic diagram of R.M.C.A. medium-frequency receiver model IP-501-A.

or low frequencies under conditions of interference, it will be found that the selectivity is improved by throwing this switch to the *P* position. When used in the *P* or grid-bias position, the damping effect of the detector grid circuit on the preceding tuned circuit is reduced and the selectivity thereby improved.

g. Calibration Curves.—Typical curves showing the dial positions for the various coils are supplied with each receiver and should prove useful as time savers when searching for signals in any particular band.

279. Medium-frequency Receiver. *a. Description.*—One of the best known receivers in this class is the model IP-501-A receiver supplied by the Radiomarine Corporation for use on board ship. This receiver is an outgrowth of the famous model SE-143 receiver designed by the U.S. Navy during the World War. Although these receivers are now used altogether with a vacuum-tube detector, provision is made through a two-way switch for the use of a crystal detector if necessary. The wave-length range of this receiver is from 250 to 8,000 meters, which is

equivalent to the frequency band from 37.5 to 1,200 kilocycles. This range may be extended, as will be explained in the next section. A schematic diagram of this receiver is shown in Fig. 289, a photograph in Fig. 290, and a wiring diagram in Fig. 291.

Theoretical Operation.—When the transmitting key circuit is opened and the proper relay contacts are closed, the transmitting antenna will be connected directly to the primary winding of the receiver.

An incoming signal will then pass through the primary winding of the tuner through the antenna series condenser and to the ground. This complete circuit then oscillates at a definite radio frequency depending upon the adjustments of the antenna inductances and the series condenser.

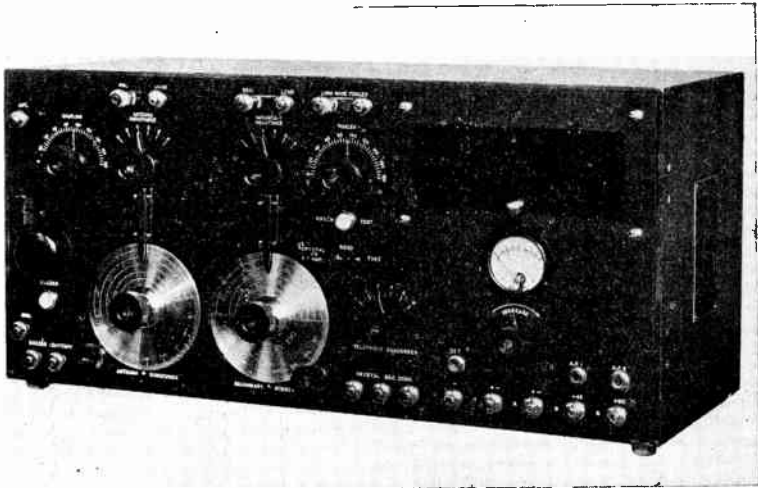


FIG. 290.—Front view, R.M.C.A. receiver, model IP-501-A. (Courtesy of R.M.C.A.)

The varying magnetic field which will be present in the primary winding L_1 under these conditions will induce an e.m.f. across the secondary winding L_2 of the tuner and across the tuning condenser C_1 . The inductance L_2 and C_1 together with the loading inductance forms the complete secondary tuning circuit.

When this circuit is carefully adjusted to resonance with the incoming signal by varying the inductance and the capacity, a maximum signal e.m.f. will be applied to the grid of the detector tube through the series-grid condenser. In other words, the signal voltage variations across L_2 and C_1 will be applied between the grid and the filament of the detector tube which will vary the steady plate current flowing in the plate circuit and the primary winding of the first audio transformer. The high-frequency inaudible currents will be by-passed through the radio-frequency by-pass condenser C_2 and the rectified currents will pass through the primary winding of the audio-frequency transformer.

The grid-condenser and grid-leak action affects the rectification of the signal so that it may be heard ultimately in the telephones.

The rectified signal variations in the primary winding of the first audio transformer are then induced into its secondary winding and into the grid of the first audio tube. Here the signal variations are again amplified in the plate circuit of the tube and transferred to the second audio stage. The highly amplified variations now

present in the plate circuit of the last tube pass through the telephone windings in which the current and magnetic variations actuate the diaphragm which render the signals to audibility.

Regeneration and oscillation are obtained by a variometer connected in the plate circuit, in which a portion of its inductance is magnetically coupled to the secondary circuit. An oscillation test button is provided for determining if the receiver is oscillating properly for the reception of continuous waves (CW). When this button is closed, the circuit will stop oscillating causing a click to be heard in the telephones. If no click is heard, this indicates that the variometer has not been adjusted for oscillation. For CW reception the variometer should be varied until an oscillation click is heard in the telephones.

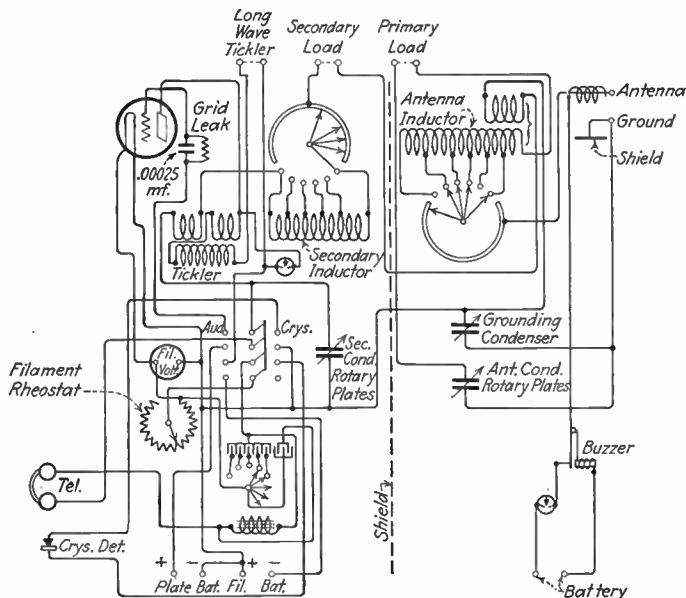


FIG. 291.—Wiring diagram of the model IP-501-A receiver.

b. Practical Operation.—In using these receivers with a crystal detector¹ throw the cam switch (see Fig. 291) to the right. Connect the crystal detector stand to the terminals marked accordingly. Connect the telephones to their respective binding posts. The buzzer should then be placed in operation by pressing the buzzer button. Adjust the crystal to a point of maximum sensitivity. Place the *coupling knob* at 180 degrees on the dial. Make sure that the antenna and ground connections are made, and also that the small safety gap across the antenna binding posts is not short-circuiting the posts. Place the dials of the condensers at zero and commence increasing the antenna and secondary inductance switches to the inductance approxi-

¹ *Note on Crystal Detector:* The crystal detector which is now obsolete depended for its operation on the unidirectional-conductivity characteristics of certain crystals (the most common of which were silicon, galena, and carborundum) when contact was made between the crystal and a contact point, wire, or another crystal. Rectification of the radio wave train was thus accomplished which ultimately resulted in pulses of direct current through the headphones and an audible note. See the authors' "Radio Operating Questions and Answers," for a more detailed explanation of this type of detector.

mating the desired wave. The secondary and antenna condensers are then varied for a sharper signal. If it is found that the tuning is broad, however, readjust the *coupling knob* to about 100 degrees and then again carefully adjust with the secondary and antenna condensers. This is simply a matter of experience, and the operator will find these tuners extremely simple to operate after a few days of use. In receiving wave lengths below 8,000 meters, the three pairs of binding posts on the top of the panel marked *LOAD* should be short-circuited. These posts are only used when the receivers are accompanied by a standard long-wave attachment placed on top of the IP-501-A tuner.

When, as is usual, the vacuum-tube detector is to be used, the cam switch should be thrown to the left. Turn the rheostat handle in the direction marked *INCREASE FILAMENT CURRENT* until the filament voltmeter reads 5 volts. If the tuner is now to be adjusted for spark reception the *TICKLER* knob should be set at approximately 120 degrees which prevents regeneration.

Vary the antenna and secondary inductances and condensers until a maximum signal is heard. Then resonate for a maximum signal by carefully adjusting the secondary and antenna condensers critically. Adjust the telephone condenser knob until the signal is maximum. Then, for sharp tuning, loosen the coupling by varying the coupling knob to approximately 100 degrees; then carefully resonate the antenna and secondary circuits again by the vernier adjustments on both aerial and secondary condensers.

For the reception of continuous waves, it will be necessary to loosen the tickler coupling as much as possible. Loosen the coupling between the primary and secondary circuits by the coupling knob and then tune the primary and secondary circuits in the same manner as before. It may be suggested here that, when the tube is oscillating, a slight click will be heard as the condenser passes through resonance. The best tone will be obtained when the setting is slightly above or below the resonant point.

Oscillation may be determined by pressing the button marked *OSCILLATION TEST*, and if a click is heard the receiver is oscillating. If the circuits are resonated and no click is heard, then vary the tickler knob either way, depending upon the station, until the desired effect is obtained.

c. The Antenna Required.—As this set is not provided with radio-frequency amplification, it can work properly only when used with the conventional ship antenna for which it was designed.

280. Long-wave Attachment.—The wave length of the model IP-501-A receiver may be extended to 20,000 meters (15 kc./s.) by the addition of the long-wave loading unit type IP-503 which is contained in a separate cabinet which is set on top of the receiver as shown in the photograph (Fig. 274).

The long-wave unit has six binding posts suitably marked which are the terminals of loading inductances for the primary, secondary, and tickler circuits of the receiver. A wiring diagram of the long-wave unit is shown in Fig. 292. The six terminals for the primary, secondary, and tickler circuits are so placed that when the cabinet is placed on top of the receiver these binding posts line up opposite the binding posts on the receiver to which they are to be connected. This facilitates the addition of the long-wave unit to the receiver proper. In addition to these connections, the antenna connects to the long-wave unit, to the

terminal post marked **ANTENNA**, and the respective connections from the antenna and ground-terminal posts on the receiver to the proper posts as marked on the long-wave unit must be made.

A rotary switch (*S*), the knob of which is situated in the lower center of the panel of the long-wave unit, provides for the selection of the desired loading inductances. When this switch is in the **LOW** position, the long-wave unit is out of the circuit and the receiver alone is worked as usual. When it is in the **MEDIUM** position, the primary and tickler loading-inductance circuits in full, and a part of the secondary loading inductance, are added to these respective circuits of the tuner (receiver). When the switch is in the **HIGH** position the primary, secondary, and tickler-loading inductances in full are connected in their respective circuits on the tuner.

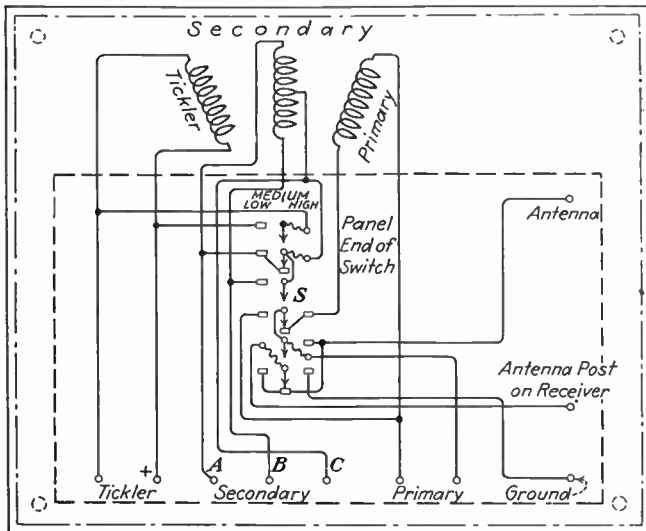


FIG. 292.—Wiring diagram of the R.M.C.A. long-wave attachment.

281. Common Receiver Troubles Summarized.—The troubles listed below make their remedies apparent, once the trouble is located.

TABLE IX.—TROUBLES IN RADIO RECEIVERS

Headphone noises:

1. Loose connections in the circuit or in phone-cord tips.
2. Defective grid leak.
3. Corroded or loose contacts of the tube prongs or sockets.
4. Broken-down grid condenser.
5. Defective tube.
6. Abnormal filament current.
7. Power-line induction.
8. Defective or run-down A or B batteries.
9. Loose antenna or ground connections.

TABLE IX.—TROUBLES IN RADIO RECEIVERS.—(Continued)

10. Poorly soldered connections.
11. Excessive plate or filament voltage.
12. Improper regeneration adjustment.

No signals:

1. Open grid circuit in the detector or amplifier circuits.
2. Tuning condenser short-circuited.
3. Poor contacts on inductance switch.
4. Open tuning coils.
5. Improper value of grid-leak resistance.
6. Open plate circuit.
7. Improper filament voltage (weak A battery).
8. Deactivated filament or cathode.
9. Broken antenna or ground connections.
10. Defective headphones.
11. Tickler connections reversed (no CW signals can be received).
12. B batteries run down.
13. Broken connections (general).
14. Poor telephone plug and jack contacts.
15. Grounded antenna.
16. High-resistance connection due to a poorly soldered joint.
17. B battery potential reversed.
18. Short-circuited by-pass condenser.
19. Defective audio transformer due to burned-out or corroded terminal connection.
20. Improper regeneration adjustment.
21. Improper tuning adjustments.
22. Poor rheostat or potentiometer contacts.

Poor signals:

1. Poorly soldered connections.
2. Deactivated filament.
3. Open in radio-frequency transformer winding.
4. Open by-pass condenser.
5. Filament polarity reversed.
6. Improper plate-supply voltage.
7. Run-down B battery.
8. Run-down A battery.
9. Poor ground connections.
10. Tickler reversed.
11. Poor contact between tube prongs and socket.
12. Antenna circuit open.
13. Poor antenna insulation.
14. Plate potential too high (results in distorted spark signals).
15. Improper tuning adjustments.
16. Improper regeneration adjustment.

III. AIRCRAFT AND POLICE RECEIVERS

282. Aircraft Receivers: General Considerations.—The design of aircraft receivers calls for important departures from conventional receiver design. Many features not required on ground-station and

marine-type receivers are highly desirable on receivers designed for installation aboard aircraft.

The aircraft receiver must combine high sensitivity with a minimum number of tubes, and it must of necessity operate on a small power supply. Both of these requirements are met by the use of screen-grid and single-grid high μ tubes. The aircraft receiver must be easy to adjust and be positive in its adjustment with a minimum number of controls. It may or may not be designed for operation on the long- and short-wave aviation-service bands.

The aircraft receiver must be light in weight, and it is desirable that automatic volume control be provided in addition to manual control. Finally, the aircraft receiver must be as immune as possible from the effects of weather, especially in so far as it affects its frequency calibration. The aircraft receiver must also be thoroughly shielded to reduce induced interference from the ignition system to a minimum.

283. Medium-high-frequency Aircraft-communications Receiver: W.E. 12A Receiver. *a. Description.*—The W.E. 12A receiver may be arranged for either local- or remote-control operation. It covers the frequency bands of 2,750 to 4,300 kc./s. and 4,300 to 6,500 kc./s., with a suitable overlap. It may be adjusted to any one frequency in each of the two bands, the desired frequency being selected by means of a mechanical control which may be extended to a remote point.

The beating oscillator in this receiver is designed for crystal-control operation in order to insure that the receiver is always adjusted to the proper frequency. It is possible, however, to dispense with the use of the quartz crystals and to employ a self-excited oscillator circuit. The self-excited oscillator is made available by the use of a specially designed, tuned circuit which is inserted in the grid circuit of the oscillator tube, thus converting it to a tuned-grid type of oscillator. When the self-excited oscillator is used, an external vernier control is required to compensate for small variations in the oscillator frequency.

The plate-supply voltage for this receiver may be obtained from a dynamotor operated by a 12-volt storage battery while the filament supply may be obtained directly from the same 12-volt battery. The plate-circuit drain is approximately 40 ma. at 200 volts, while the filament drain is approximately 3.2 amp. at 12 volts. The heater current for each of the quartz plates used in the crystal-oscillator circuit is approximately 1.2 amp. at 12 volts; this may also be supplied by the storage battery.

The W.E. 12A receiver is designed for use with headsets only. Adequate output is provided to operate two pairs of low-impedance headsets connected in parallel. This receiver employs eight vacuum tubes as follows:

VACUUM TUBES W.E. 12A RECEIVER

All tubes are W.E. 283A, except V_8 which is a W.E. 285A

V_1 Radio-frequency amplifier.

V_2 Modulator (first detector).

V_3 Beating oscillator.

V_4 First intermediate amplifier.

V_5 Second intermediate amplifier.

V_6 Third intermediate amplifier.

V_7 (second) detector and AVC.

V_8 Audio amplifier (pentode).

The circuit employed and other important features of the receiver will become apparent from a study of the schematic wiring diagram given in Fig. 293.

b. Operation.—The intermediate-frequency amplifier which contains three stages of highly selective amplification operates at a frequency of 385 kc./s. Four intermediate-frequency filters are used as coupling elements between the modulator, the three intermediate-frequency stages, and the detector circuit. Each of these filters contains two tuned circuits which are inductively coupled to constitute a band-pass filter having a band width of approximately 8 kc./s. with a mid-band frequency of 385 kc./s.

The detector circuit employs a two-element rectifier which is used as a combined detector tube and automatic volume control. While a screen-grid tube is used, the control grid and cathode are the only active elements, as both plate and shield grid are connected to ground. Resistance-capacity coupling is employed between the detector and the audio-frequency tube.

A pentode tube is employed in the single-stage audio-frequency amplifier. When the transmitter is in operation, the output tube functions as an amplifier to raise the side tone to the desired level. The operation of control relay S_3 changes the proper connections when the transmitter is placed in operation.

Automatic gain control is obtained by permitting the incoming signal to readjust the grid bias of the radio-frequency amplifier, the modulator, the three intermediate-frequency (*IF*) amplifiers and the audio-frequency amplifier by an amount proportional to the strength of the received carrier. When a signal is applied to the grid of the detector tube, a rectified voltage which is proportional to the carrier amplitude appears across a resistance in the cathode circuit of the detector. This voltage is of the proper polarity to increase the negative bias as the strength of the incoming signal increases. This additional bias reduces the gain of the entire receiving system and tends to hold the audio-frequency output at a constant value regardless of the strength of the received signal.

The level control and the manual gain control are not a part of the W.E. 12A receiver but are a part of the control unit which will usually be located at some distance from the radio receiver in a location convenient to the pilot or radio operator.

The level control is a potentiometer connected in the output circuit of the receiver, by means of which the output level may be varied to bring the signal to the desired loudness. The control is arranged so that it does not affect the output level of any other receiver, such as a weather and beacon receiver, which may be connected to the headset in parallel with the output of the W.E. 12A receiver.

As stated before, the receiver is primarily fitted with a crystal oscillator, although a self-excited oscillator may be used to produce the beat frequency. While the self-excited oscillator will prove entirely satisfactory under practically all working conditions, the use of the crystal-controlled oscillator is recommended whenever extreme stability is required and wherever it is objectionable to make the necessary adjust-

ments of the vernier control. In addition, the use of crystal oscillators is strongly recommended wherever large temperature or humidity variations are frequently encountered.

The same antenna will generally be used for the receiver as is used for the transmitter, when the receiver is used as part of an aircraft-communication system. If the receiver is used alone and is not a part of a two-way system, a horizontal or inclined antenna, running from a mast mounted on the forward part of the fuselage to the tail of the plane, may be used. While the location of the antenna depends upon the structure of the ship, the height of the mast should be 6 to 8 ft. if the available hangar clearance permits. The receiver should be mounted as close to the antenna lead in as is practicable so as to avoid long horizontal connecting wires between the antenna and the receiver.

The receiver has a tuned input circuit which may be used with any antenna having a capacity of approximately 150 mf. or larger. The antenna must not be resonant, however, at any frequency below the highest frequency at which the receiver is to be used.

The antenna, modulator, and oscillator circuits are tuned by variable-air condensers which may be locked after the final tuning adjustment has been made. A modulated carrier of the desired frequency must be used to make the tuning adjustments. Final adjustments of the antenna tuning condensers must be made with the receiver installed in the airplane, which should be removed from the hangar while this adjustment is being made. The frequency of the crystal oscillator must be 385 kc./s. above the frequency of the signal which is to be received. Typical dial readings for the various frequency ranges of the receiver are provided in the form of calibration curves which appear in the instruction book on this receiver supplied by the manufacturer.

After the receiver has been installed and properly tuned in accordance with instruction-book directions, it is ready for operation. If the crystal-controlled oscillator is used, sufficient time should be allowed to permit the quartz plates to reach their operating temperature. This will require from 10 to 30 min., depending upon the ambient temperature. The indicator lamps on the front of the receiver will light intermittently when the crystal holders are at the proper temperature.

For any given setting of the level control the output level of the receiver will be approximately constant regardless of the strength of the received signal. The level control may be used for adjusting the volume of the signal to the desired value.

The manual gain control will be useful to reduce the amount of noise present in the headset while no signal is being received. This control should usually be turned as near the maximum position as is possible without raising the noise level to an objectionable value. The manual gain control may also be used to prevent overloading when an extremely strong signal is received. There will usually be no necessity for using the manual gain-control overloading unless the airplane in which the receiver is installed flies directly over the transmitting antenna.

284. Radio-beacon and General-purpose Aircraft Receivers.—The function of the radio-beacon receiver is primarily to receive radio-range beacon emissions and weather reports from aeronautical stations on the aviation-service frequency band from 285 to 350 kc./s. Therefore, any receiver capable of receiving these frequencies could be termed a *beacon receiver*, although its frequency range may be much wider than is required for this purpose, that is, it might also be used on the high frequencies assigned to general aviation-communication service. Such a receiver will now be described.

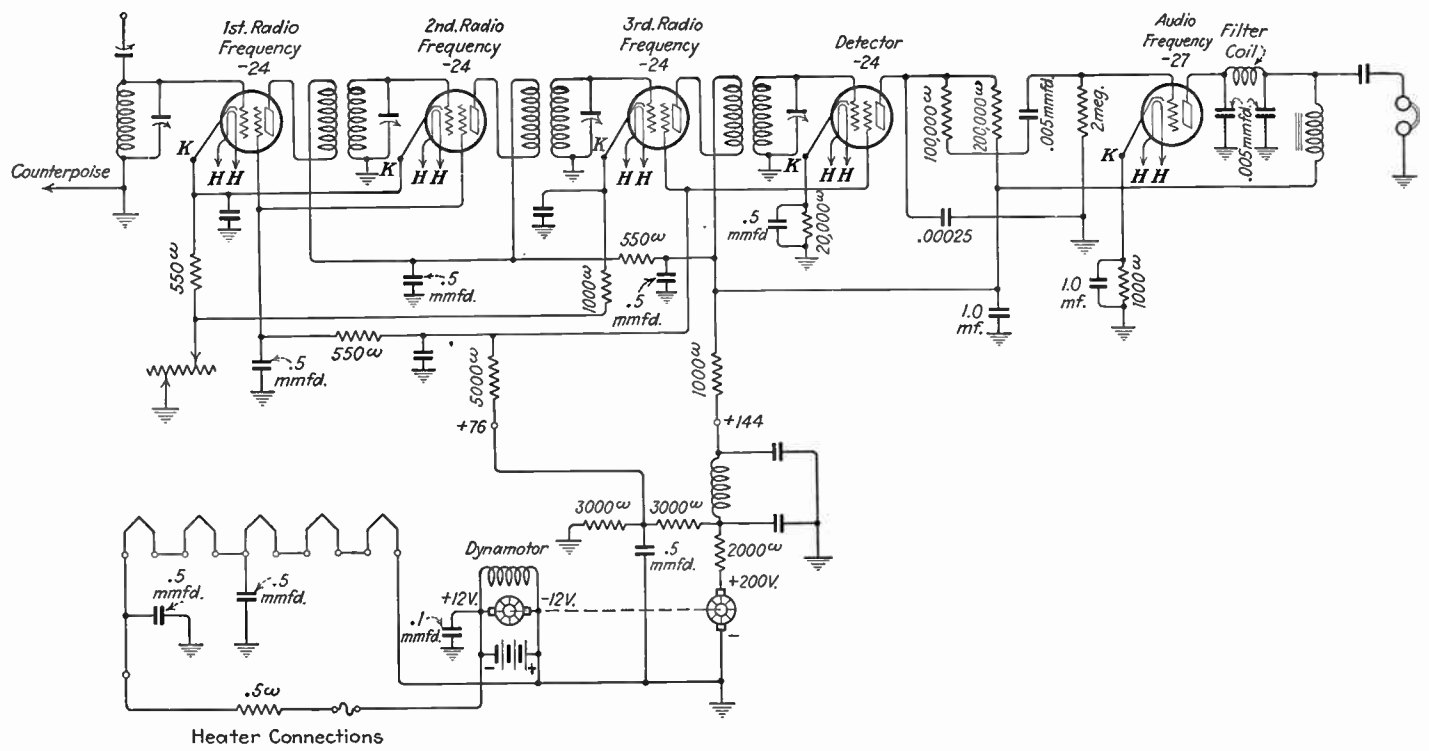


FIG. 294A.—Aircraft receiver, Stromberg-Carlson model D. (Courtesy of Stromberg-Carlson Telephone Manufacturing Company.)

Stromberg-Carlson Aircraft Receiver.—The receiver to be described is known as the model D and is designed for the reception of modulated signals on aircraft in the frequency range 235 to 8,000 kc./s. It is a tuned radio-frequency receiver having four tuned circuits and one resistance-coupled audio-amplifying stage. Coil sets are available for frequency bands as follows:

D-2023.....	235- 470 kc./s.	D-2024.....	600-1,200 kc./s.
D-2041.....	250- 500 kc./s.	D-2025.....	2,000-4,000 kc./s.
D-2040.....	500-1,000 kc./s.	D-2039.....	4,000-8,000 kc./s.

In addition to the above coil sets, a dual-coil set has been developed for use with this receiver, whereby the operator may switch from a high- to a low-frequency band without changing the coil set. A schematic circuit diagram of this receiver is shown in Fig. 294A.

Either a dynamotor supplying +144 and +76 volts or an aircraft B battery supplying equivalent potential may be used for the power supply. If a dynamotor is used, care must be taken to obtain proper low- and high-frequency filtering in the lines to the dynamotor. Improper or inadequate filtering of such a unit can reduce the effectiveness of a sensitive receiver many times. In general, dynamotor cables and dynamotor filter cases should be well grounded to the fuselage of the airplane. A radio filter choke should be used in the +12-volt line near the motor terminal. The high-voltage line should be filtered for both audio and radio. No exposed unfiltered line should be allowed to radiate back to the antenna.

If a 200-volt dynamotor is used, either the biases must be increased on the -27-type tube by increasing the auto-bias resistor to 1,500 ohms, or the 200 volts must be reduced by a dropping resistor to +144 volts.

Present-day high-grade tubes should not be microphonic. To test for this trouble, strike tubes with the fingers with a considerable blow. Good tubes will produce little noise, while poor ones will produce clicks or prolonged disturbances. The tube most sensitive to this test is the first radio-frequency tube. If a -24 tube is suspected, place it in the first radio-frequency socket and hit it rather hard. Do not be afraid to hurt the tube because if it is well constructed and in good condition no damage will be done.

Another difficulty has been the loosening of bearings in the gang condenser, in some instances. The repair of this item should be undertaken only by a careful mechanic. Trouble at this point is noticed by clicking or possible oscillation in spots over the tuning dial, and the inability to repeat results and get the effect always at the same place.

The antenna must be carefully aligned, as this receiver uses a tuned-input system to obtain a high signal-to-noise ratio. Unless this is done, best results will not be obtained.

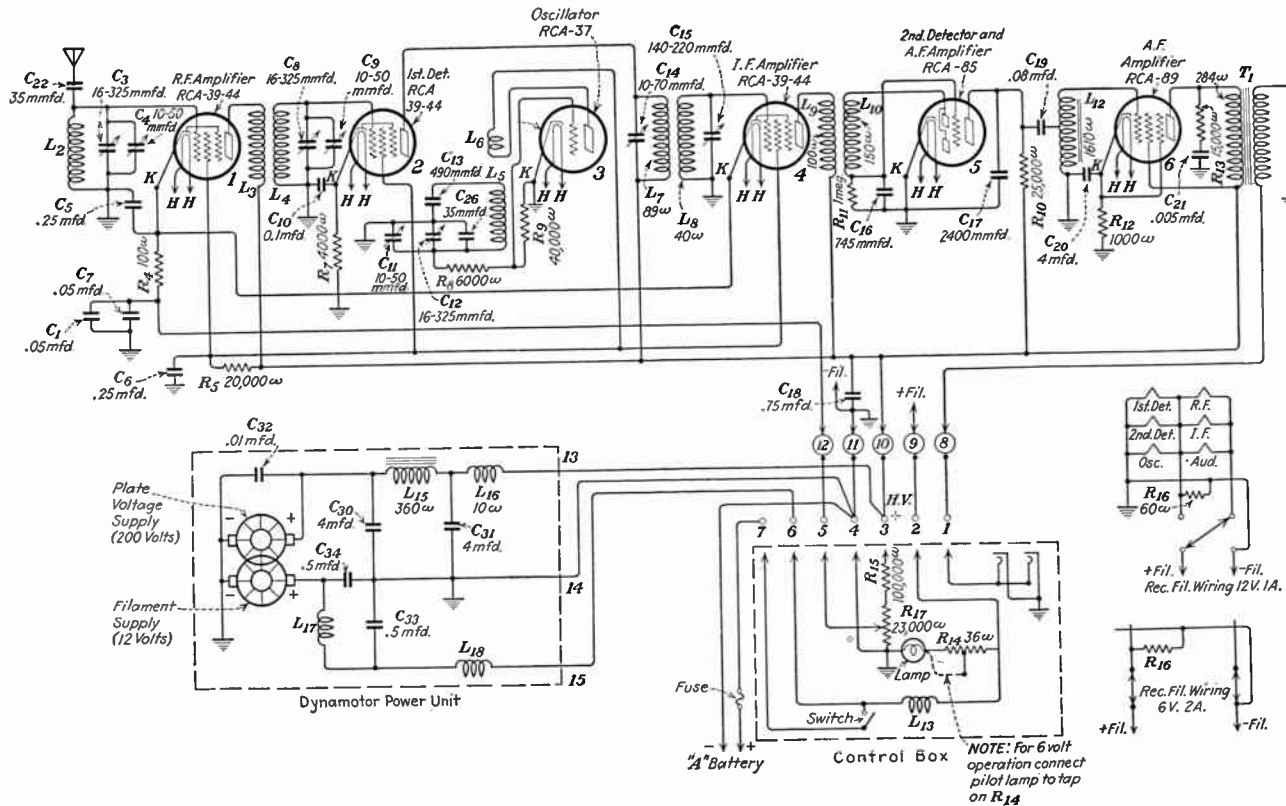


FIG. 294B.—Aircraft-beacon receiver, RCA-Victor type AVR-2. (Courtesy of RCA Manufacturing Company, Inc.)

The steel springs on the tube sockets should be inspected occasionally. A light coating of vaseline will protect them from rust. The newer sockets are fitted with plated springs and will not rust, hence they do not require the coating of vaseline. A high grade of watch oil should be used sparingly on the gang-condenser and dial mechanisms, and also on the antenna-aligning condenser.

The chassis is tightly held in place by shakeproof lock washers and screws which also make electrical contact with the supporting structure thereby providing a method of obtaining stability over a wide range of frequencies, such a method being necessary on high-gain, tuned-radio-frequency receivers.

Although these directions were formulated to apply to the Stromberg-Carlson model D receiver by the Aircraft Radio Corporation, the designers

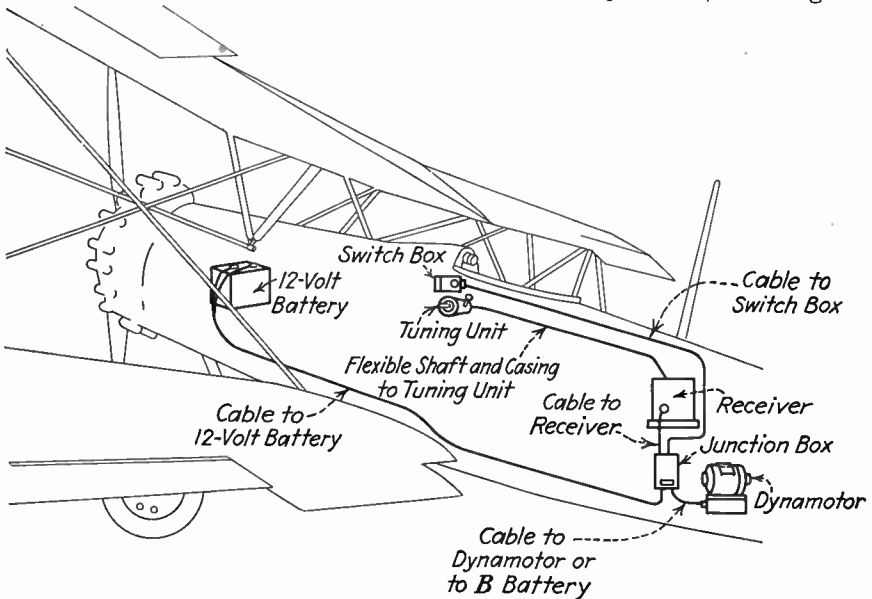


FIG. 295.—Typical method of installing radio receiving equipment in an airplane.

of the receiver, they apply in general to all similar types of aircraft receivers.

RCA Victor Aircraft-beacon Receiver.—This receiver is designed especially for radio-beacon reception and is known as the type AVR-2. It comprises a six-tube superheterodyne receiver, a dynamotor power-supply unit, and a control box for remote operation of the entire equipment. A schematic diagram of the outfit is shown in Fig. 294B.

The method of installing a receiver in the airplane is shown in Fig. 295. Note that the receiver and its high-voltage power supply are located in the rear fuselage, the entire reception being controlled by remote controls located in front of the pilot.

285. Antennas for Radio-range-beacon Reception.—Although airplane antennas have been treated from the construction angle in Chap. VI on Antennas there are certain facts of interest in connection with antennas to be used especially for radio-range-beacon reception. Each type of antenna has a distinct localizing effect, that is, each type causes the received signal to take on certain intensity characteristics while traveling near the radio-beacon transmitter. Before the development of the transmission-line antenna this localizing effect was such as to make the non-directional vertical-pole-type antenna desirable for this service.

The two principal sources of error in radio-beacon reception from loop-antenna transmitters were caused by *night effect* and *airplane effect*; the former will be explained in the next chapter, and the latter is the error introduced when the airplane is flying at an angle to the course in close proximity to the beacon station. The transmission-line antenna which is now used at radio-range-beacon stations appears to eliminate both types of errors without restricting the receiving antenna used aboard the aircraft to any specific type.

It now seems desirable to use a directional rather than a non-directional receiving antenna on airplanes. Very satisfactory results are secured by the use of a 45-degree-inclined antenna running from the fuselage back and up to the vertical fin. The same satisfactory results can be secured by using the usual vertical-pole antenna by adding a flat-top element from the top of the pole antenna back to the vertical fin.

Generally speaking a radio-beacon receiver may be used on any size antenna from 30 to 1,000 mmf. This is about the capacity achieved by the use of the usual vertical-pole or inclined antenna, before referred to. On some of the larger transport planes an antenna located inside the vertical fin of the ship has been used with success for beacon and weather-report reception.

The receiver should be mounted as close to the antenna as practical and long horizontal connecting wires between the antenna and the receiver should be avoided.

286. Aircraft Shielding and Bonding. *a. Requirements.*—The improved design of modern aircraft receiving equipment has reduced to practically a minimum degree, the problem of noisy reception due to the receiver design itself. This is primarily due to the great care exercised by manufacturers in the magnetic shielding of all radio-frequency coils and internal wiring, introduction of grid and plate filtering in all stages, automatic volume control, interstation noise suppression; tube design, mechanical rigidity, and the complete shielding of the entire receiver. If the receiver incorporates all of these improvements, particularly complete exterior shielding, it will be impossible for it to pick up any stray electrostatic or magnetic fields produced by exterior ignition or contact interruption due to switches, loose connections, high-frequency

contact variation, or any other source of impulse excitation. However, when the antenna or loop system is connected to the receiver, all such electrical disturbances will be picked up by the antenna or loop and amplified through the receiver.

The problem, therefore, with completely shielded receivers, is the elimination of noise interference at the source rather than in the receiver itself. Hence, great care must be exercised to reduce electromagnetic radiation from all undesirable sources by thorough shielding of all ignition wires, spark plugs, spark coils, lighting wiring, electrical indicating devices and associated wiring, good electrical connections, and by thorough grounding of all shields to a common ground connection. The importance of connecting all metal parts to a common ground is extremely vital since any variation between metal parts, due to a high-resistance contact or connection, will produce a potential difference between the two or more contacts which may result in electrostatic or electromagnetic radiations which are picked up by the antenna or loop system and amplified.

b. Shielding.—The principal function of electrical shielding in receiving circuits, as previously pointed out, is to prevent inter-circuit magnetic and electrostatic reactions and the picking up of highly damped electrical surges or impulses. In aircraft (and automobile) radio systems the latter is of the most vital importance since it is the source of external electrical disturbances. Consequently, if the radiation of highly damped electromagnetic radiations are suppressed at the source, interference due to these may be practically eliminated. These radiations in aircraft (and automobiles) are primarily due to the spark-ignition system and its associated electrical connections, the generator circuits and the lighting circuits. At any point in which electrical circuits are broken because of the opening of switches, the production of sparks, and the variation of magnetic fields, such as in the generator commutator, highly damped waves are radiated into space or along conducting wires. Hence, all radiations must be suppressed or limited to the confines of the sources creating them. This may be accomplished as follows.

The high-tension system including either the magneto or coil systems should be placed into a complete metal housing. The leads should be enclosed in a continuous metal sheath extending from the housing to their point of termination. Each individual lead for the spark plugs should be encased in a flexible copper-braided shield, the braided shield to be soldered to the manifold all around the holes through which the wires pass. If the leads emerge from an insulating block, this block should be covered with a metal shield or can. All braids must be soldered to the can around the hole through which the wire emerges. All braiding or piping must make contact with the engine metal and chassis at as many points as possible. These contacts should be made by short lengths of

pigtail braiding or copper stripping thoroughly soldered to the points of connection.

All spark plugs must be enclosed in a properly designed shield housing possessing a complete metallic circuit, each shield to be soldered to a copper braid or strip and grounded to the engine frame and chassis.

All leads in the low-tension system including the light wiring, instrument-board connections, and generator must be similarly shielded and grounded to the engine frame and chassis. The instrument board and battery should be thoroughly encased in a copper housing and grounded.

There are available complete ignition harnesses enclosed in flexible braiding. These should be additionally encased in a pipe manifold to increase mechanical rigidity and protection; both harness and piping to be thoroughly grounded.

Summarizing, it may readily be seen that the thorough elimination of radiation and pick-up is almost entirely dependent upon complete shielding and frequent grounding of all wires and component parts. If the entire engine and all parts are thus thoroughly shielded and grounded, complete interference elimination may be approached.

c. Bonding.—This term refers to the interconnection of all metal parts in the aircraft to a common point or ground (chassis), in such a manner as to provide a path of extremely low resistance to the flow of any high-frequency currents which may be developed by the accumulation of electrostatic charges. If charges are accumulated between two or more metals, because of resistance between them, the electric field produced will be stationary and no radiation will result. If, however the resistance between these metals is varied owing to vibration or rubbing, the friction will produce a variation in the contact resistance which will produce a movement of the electrostatic field resulting in the radiation of a highly damped (broad) wave or pulse of high frequency. It is possible that the electrostatic accumulations between metals due to poor contact may develop an electrostatic field or e.m.f. sufficiently high to cause a spark to jump between the contacts or to an adjacent conductor. This will result not only in severe pulse radiation but will also introduce a serious fire hazard.

In wooden airplanes, therefore, it is necessary to provide metal strips of copper about $\frac{1}{4}$ by $\frac{1}{64}$ in., or larger along the front and rear spars from the fuselage to the wing tips. These strips must be connected from the engine base or metal chassis to the internal brace wires at each drag brace, to the external brace wires and fittings wherever the wires and fittings come into the wing, and to each other by similar strips which run from the front and rear strips along each drag brace. The front and rear strips must also be connected together at the wing tip. All brace wires or any other movable or semimovable parts which come into actual electrical contact with each other, or with other metal parts of the ship,

must be thoroughly connected. Where wires are very close but no actual contact exists, insulation must be provided to prevent these parts from touching because of vibration.

Turnbuckles and tie-rod terminals should be bridged by a flexible copper braid and grounded to prevent high-resistance contacts from developing between them. Usually these contacts are quite secure and introduce practically no resistance, in which circumstances no shunting braid is required. These connections should be tested frequently, however, owing to the possibility of high-resistance formation from moisture and corrosion. All water, gasoline, and oil lines must be grounded to the metal chassis at intervals. Rubber-hose joints must be bridged by copper-braid strips to insure against the metal pipes on either end of the hose being at different electrical potentials. All metal hinges, control wires, elevators, ailerons, bolts, and screws must be grounded to the chassis. In metal planes bonding is unnecessary, but all inter-connecting plates, control wires, bolts, and electrical wiring metal cables must be connected by copper braids or welding. If metal parts are separated for shellacing, they must be carefully scraped and cleaned when reassembling.

The thoroughness of the bonding and grounding between various metals and the chassis can be determined by the use of an ohmmeter and a battery. All metal parts should indicate a full-scale reading (0 ohms), when the meter is connected across the chassis and the individual parts under test. If long test leads are used with the meter, an exact value of the lead resistance must first be determined to avoid erroneous readings. A small flashlight test lamp or buzzer may also be used by connecting either one of them in series with a small battery to one side of the metal under test and the other side to the chassis. If these tests are made, while the motor is running to produce vibration, an accurate indication of the efficiency of the bonding may be ascertained. If the lamp or buzzer is intermittent in its operation it is quite evident that the grounding is poor. Care should be exercised that no sparks are produced when using the lamp or buzzer. It is advisable to use an ohmmeter when making tests for proper grounding when the motor is in operation.

The component parts which must be bonded may thus be summarized as follows:

1. All metal tubing and cables from the ignition system.
2. All connecting wires leading to the radio-receiver instrument panels, electrical lighting.
3. All spark-plug caps and shielding cans.

All of these parts *must* be bonded to a common point through a low-resistance conducting strip or braid to avoid potential gradients between various grounded portions. If possible, the bonding should be made in even sequence at every 7 or 8 in.

287. Police Radio Receivers. *a. Precinct Receiver.*—An example of this type of receiver made by the United American Bosch Corporation is of the superheterodyne type and employs four tuned intermediate circuits at a frequency of 456 kc./s. A schematic circuit diagram of this receiver is shown in Fig. 296. This receiver is of the fixed frequency type; that is, it is adjusted to the desired frequency and the adjustments locked in position so that the receiver will not accidentally become detuned.

The preselector is of the high-gain type which assists materially in reducing the signal-to-noise ratio. The output of the preselector is fed to a type 58 tube as the first detector where the voltage is injected from the oscillator which is a type 56 tube. The output of the first detector is

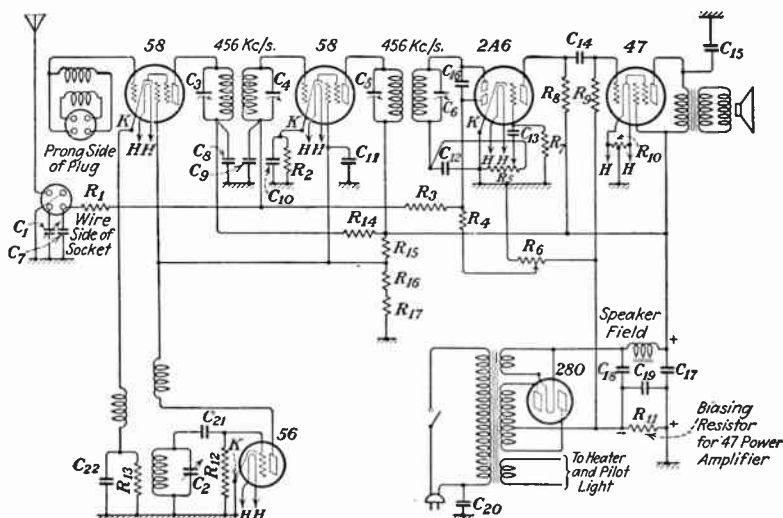


FIG. 296.—Police precinct receiver. (Courtesy of United American Bosch Corporation.)

fed through a double-tuned intermediate transformer to the amplifier tube, which is also a type 58. The output of the amplifier is fed through another double-tuned stage to the second detector, which is a double-diode, high- μ triode, type 2A6 tube. One of the diodes serves to develop audio-frequency energy which is fed by the volume control to the input of the triode section which is the first audio stage. This triode is resistance coupled to the power stage which is a type 47 tube. The other diode in the 2A6 tube serves to develop voltage for automatic control of the gain in the preceding amplifiers.

There is also provided a sensitivity control which can be adjusted to govern the maximum sensitivity which the automatic control will allow the receiver to assume when there is no carrier present. In locations where there is a strong signal, this can be set nearly to minimum sensitivity thereby eliminating the greater portion of between-transmission noises.

In localities where the signal is weak it is obvious that the sensitivity control must be advanced toward the maximum position to insure the signal being received.

An antenna of 30 to 40 ft. in length will generally be found sufficient where the signal is strong. Where the noise level is high, the antenna can be placed at a remote point away from the noise source, at right angles to power lines, and a shielded transmission line run from the antenna to the receiver. It will be necessary to use a step-down transformer at the antenna, but none is necessary at the receiver as this is already adapted to a low-impedance input. In localities where the signal is weak, a longer antenna of 60 to 75 ft. will increase the signal pick-up.

A ground connection is not absolutely necessary and, in some installations, may cause an increase in the noise. Noise may often be caused by using a ground-lead return which is common to other equipment such as bells and teletypes. If possible, a ground which is entirely independent of other equipment is desirable.

b. Police-car Receivers.—A widely used police-car receiver of the fixed-tuned type, made by the United American Bosch Corporation, is shown in Fig. 297 and is described by the manufacturers as follows:

This receiver can be used as a seven-tube receiver with class B audio amplification and dynamic speaker or as a five-tube receiver with magnetic loud-speaker. The use of five tubes and a magnetic speaker effects a considerable decrease in A-battery drain.

This receiver is of the tuned radio-frequency type and has automatic volume control accomplished by a duplex diode-triode detector. The B power supply may be furnished by B batteries or a magmotor.¹

The set has one of two numbers determined by the tuning range. It has fixed tuning for each of the police broadcast bands, although it is adjustable to cover a range of frequencies on each side of these bands. The model 113-X is designed for the low-frequency band (1,500 to 1,800 kc./s.) and the model 123-X for the higher frequency band (2,250 to 2,500 kc./s.). The receiver can also be furnished for other frequencies, but is supplied correctly adjusted to any specified frequency.

The receiver utilizes either a capacitor plate underneath the car or a conventional roof antenna, or an easily installed copper-screen antenna. A shielded lead-in cable is provided for the connection between antenna and receiver.

Either of two types of loud-speakers may be used. The magnetic type is particularly adapted for installation on the roof above the driver's seat on closed cars. The dynamic type employs a permanent magnet field and consequently draws no current from the battery. Both speakers are provided with mounting brackets, and plug connections to the chassis. When the magnetic speaker is used, remove both type 52 tubes and operate the set as a 5-tube receiver—no other changes are necessary.

¹ Manufactured by the United American Bosch Corporation, Springfield, Mass.

The receiver draws the filament or A current directly from the battery, and the B voltages from the magmotor, which is also connected to the battery. B batteries may also be used.

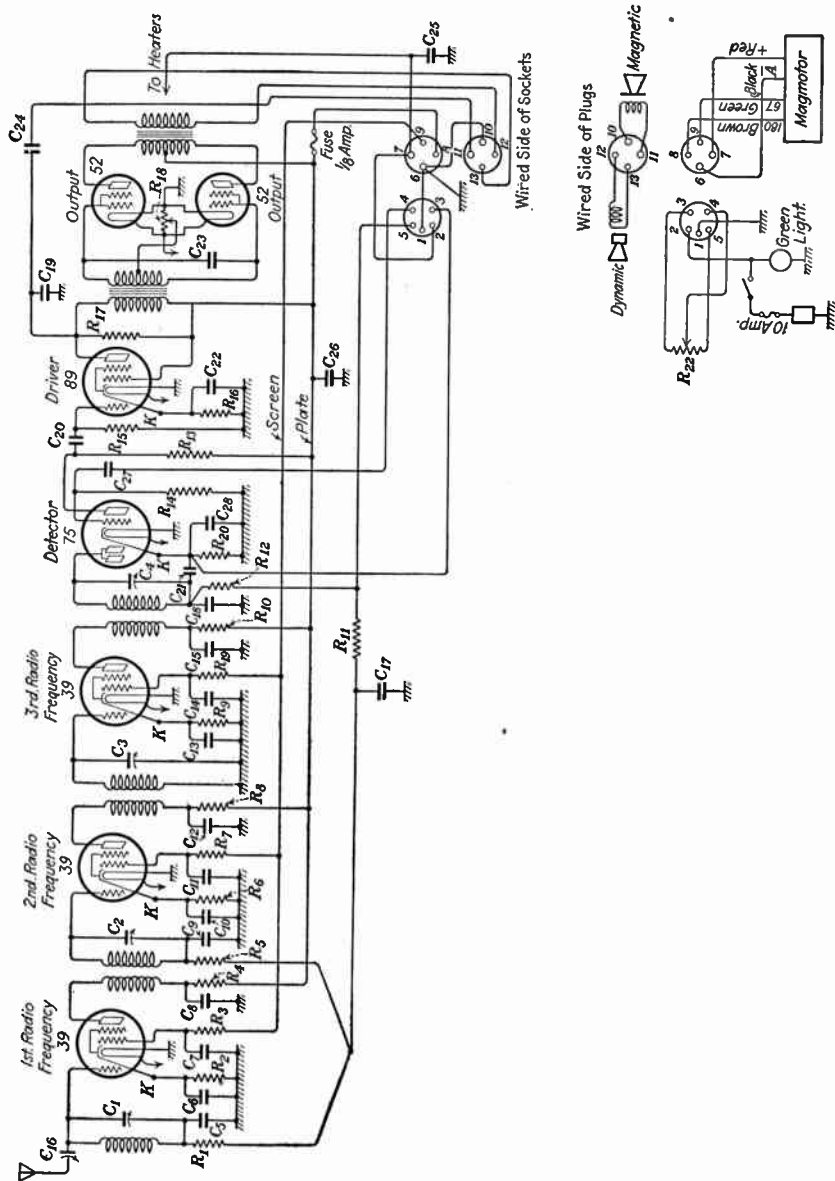


Fig. 297.—Police car receiver. (Courtesy of United American Bosch Corporation.)

The power connection is made through a single wire to the high side of the car battery. The return circuit is through the frame. The polarity of the magmotor is adjusted for each installation and care must be taken not to reverse the battery polarity if the battery is removed for any reason.

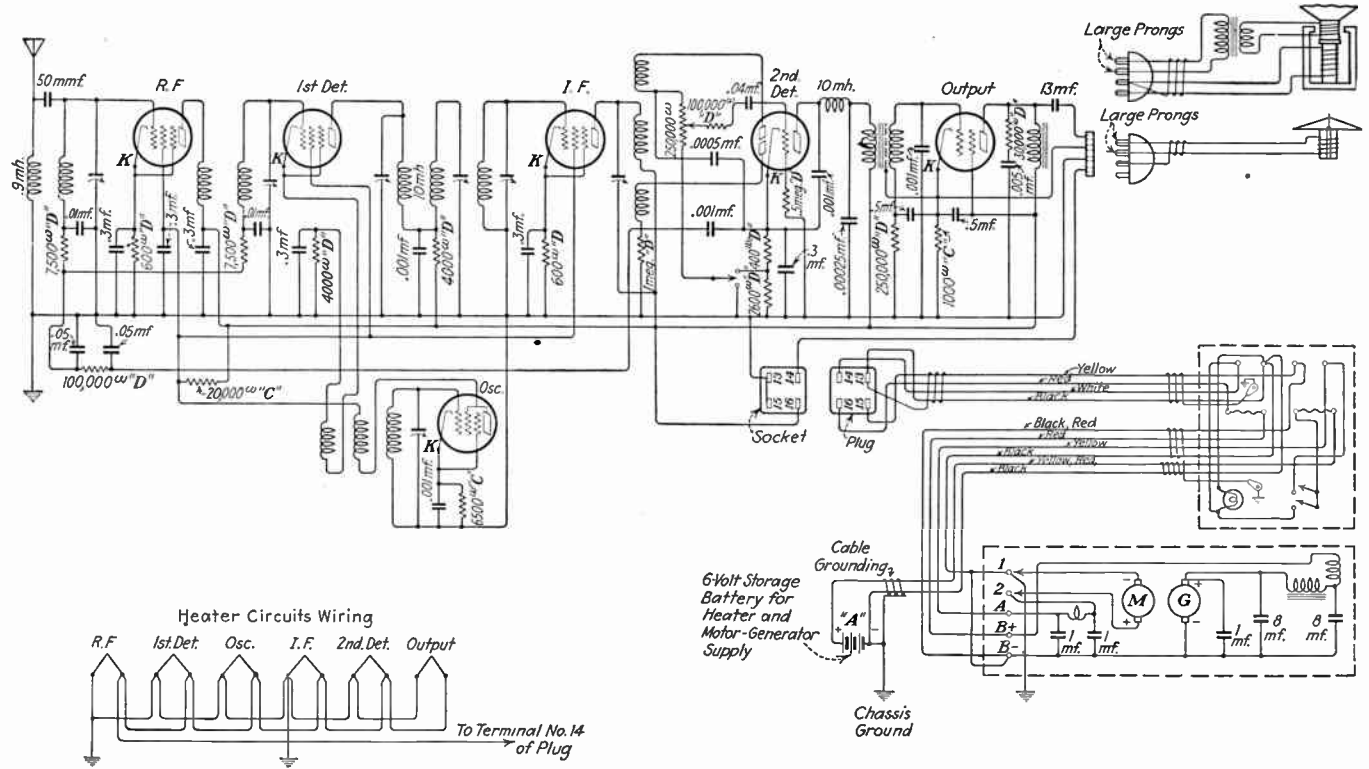


FIG. 298.—Police car or precinct receiver. (Courtesy of Stromberg-Carlson Telephone Manufacturing Company.)

The magmotor is essentially a dynamotor having a double commutator armature which has two separate independent windings, one wound for 6 volts as a motor winding, and the other wound for 135 volts as the generator winding. Both of these windings occupy the same armature slots and thus form a very compact and efficient unit since only two sets of bearings are required. The 6-volt winding is connected to a seven-bar commutator which is supplied with current from two copper-graphite brushes. The high-voltage winding is connected to a 14-bar commutator which delivers current to two high-resistance carbon brushes. The two-pole field structure is magnetically excited by means of an ordinary horseshoe magnet, very similar to that used on magnetos. This has the advantage of compactness and higher efficiency since the field current is eliminated.

The complete magmotor assembly is mounted on a steel base with four brackets which support the cover and provide the means for mounting the unit on the car. The power unit is insulated from its mounting by means of rubber cushions which permits the motor to run freely and quietly, completely eliminating vibration and noise. All connections from the magmotor are combined in a single cable terminating in a plug which connects with a socket in the receiver.

Another widely used police-car receiver is the Stromberg-Carlson No. 31. This receiver utilizes the superheterodyne circuit of the fixed-tuned type which is shown in Fig. 298. It is made for mounting behind the dashboard or the rear of the front seat of a sedan.

Tuning adjustments are designed so that they cannot become changed accidentally and are accessible only when the receiver is removed from its mounting plate. The antenna lead consists of a 6-ft. length of shielded high-tension cable connected to the receiver by a separable plug.

The loud-speaker is a specially designed magnetic cone-type unit contained in an aluminum housing and designed for mounting either on the ceiling or in one of the rear corners of the car. It connects to the receiver by a 10-ft. length of cord and a separable plug.

The receiver is highly sensitive, giving an output signal of about 50 mw. from an input of $0.25 \mu\text{v}$. It has a maximum power output of approximately 800 mw., sufficient to produce an intelligible signal under the most severe noise conditions that could be encountered.

The total drain on the car battery with this installation is 4.8 amp.; of this 1.8 amp. go to the filaments.

The No. 31 receiver chassis is also supplied mounted in a cabinet together with a loud-speaker and power supply for use as a precinct receiver (D-2555). A hole in the side of the cabinet permits access to the volume control.

288. Shielding on Automobiles.—The fundamental methods of shielding, as outlined for aircraft in Sec. 286 on Aircraft Shielding and Bonding, apply equally well to the automobile. In addition, certain

supplementary methods of noise suppression may be used on automobiles in instances where it is not practical to follow out the thorough methods which are standard practice for aircraft. Using these supplementary methods it is frequently possible to reduce interference to a minimum by the use of suppressors and filters and the proper arrangement of the antenna system.

This is accomplished by the use of suppressor resistances or radio-frequency choke coils and by-passing condensers. The use of a high value of resistance (approximately 25,000 ohms) in series with each of the high-tension wires leading to the spark plugs has been used with fair success. This method, however, is inefficient since it affects the efficiency of the engine performance by reducing the spark intensity due to the voltage drop across each resistance when the sparking current is passing through them. This defect may be overcome by the use of a larger ignition coil to compensate for the voltage drop. However, since the objective is to suppress the amplitude of the high-frequency surges passing through the high-tension wires by the introduction of a loss, the desired condition can be brought about much more efficiently by the use of a radio-frequency choke coil in series with each high-tension lead. A coil wound with No. 30 insulated iron wire with an inductance of approximately 5 mh. will have an impedance value to the high-frequency surges in excess of 25,000 ohms and yet will possess a very low direct-current value of resistance. With this arrangement, effective noise suppression will result without the usual drop in ignition voltage associated with the high-resistance method.

The following procedure is suggested:

1. With the engine stopped, install the suppressors on the spark plugs. Start the engine and, with the receiver turned on, observe the amount of interference.

2. With the engine stopped, install the suppressor in the center high-tension cable as close to the distributor as possible. Never install this or any other suppressor in the cable at the coil. Start the engine and, with the receiver turned on, observe the amount of interference.

3. A 1-mf. condenser connected across the commutator terminals of the generator will reduce noise suppression through the absorption of the commutator ripples and sparking. Similar condensers should be connected in the low-tension circuits such as in the light and dashboard wiring circuits, and across any device producing interference; these condensers to be placed between the ungrounded wires and the ground. These condensers serve as by-passing devices for radio-frequency surges picked up in the wiring circuits.

Considerable interference produced by the ignition system of the automobile is picked up by the antenna itself, particularly in that portion of the antenna leading from the radio receiver to the regularly

built-in portion of the antenna on the roof top. If this lead is shielded up to the point where it enters the top horizontal portion, interference may be practically eliminated, provided of course that the shielded portion is thoroughly grounded. This method has a disadvantage in that the capacity relationship between the wire and shield, serves to reduce the signal volume due to its by-passing or shunting effect across the input circuit of the radio receiver. The capacity effect may be reduced, however, by using a concentric tubing of wide diameter with the antenna stretched tightly through the center, thereby reducing the capacity relationship. This latter method would obviously have to be included in the initial body design to make it practical.

The United American Bosch Corporation recommends the following procedure for interference elimination:

The low-tension wires to the ignition coil or ignition switch should not be run in the same tube or conduit as the high-tension wires.

Reground the grounded side of the points directly to the motor block if possible. Oil and grease in the distributor often provide a poor ground, and, as a consequence, interference travels up the spark-control wire to the dash.

Make sure that the spark gaps are set in accordance with the engine manufacturers' specification, as contained in their instruction book.

Make sure that the terminals of the high-tension cables actually make contact with the wire itself. Small sparks originating at this point cause interference.

Make sure that the spark coil is well grounded. Varnish or enamel should be removed at this point. Old or cracked high-tension wire is a source of ignition noise and poor engine performance as well. Replace with a good grade of cable.

It is sometimes necessary to use shielded high-tension cables for the high-tension wire, running from the coil to the distributor. Avoid this, if possible. Ground the shielding carefully.

Separate the low- and high-tension wires at all points, particularly where they pass through the bulkhead.

Ground the motor blocks and bulkhead together.

Be sure that the hood is closed and latched when listening for ignition disturbance.

If it is necessary to insulate the metal terminals of the suppressors use electrician's [rubber] tape. Ordinary friction tape is useless as an insulator of the high voltages used in high-tension ignition circuits.

Noise-suppression precautions in automobiles may be summarized as follows:

1. Complete magnetic shielding of all high- and low-tension cables and proper grounding.
2. Complete shielding of spark plugs and connecting wires and proper grounding.
3. Insertion of shielded radio-frequency choke coils or resistances in series with each spark-plug lead from the distributor.
4. Complete shielding of the spark coil unit and grounding.
5. By-passing of all low-tension wires to ground through a condenser.
6. Shielding of the vertical portion of the antenna lead-in to the receivers.

289. Ultra-short-wave Receivers.—This type of receiver is the latest development in the radio-receiver field and, therefore, will not attain stability of design until the very high frequencies come into extensive use. As long as there is a scarcity of transmitting stations in these spectrums, and interference is not a problem, the superregenerative circuit commonly used in this service will continue to be satisfactory. However, as transmitting stations increase, and interference results, a more selective receiver will be necessary. The superheterodyne circuit holds promise of being most suitable for ultra-high-frequency work if crystal-controlled transmitters are employed.

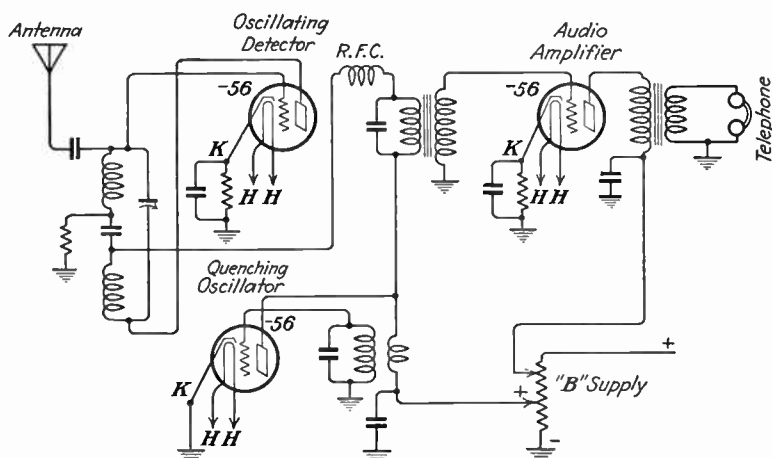


FIG. 299.—Superregenerative ultra-high-frequency receiver.

The superregenerative circuit was theoretically considered in Chap. V on Receiving Circuit Principles. A practical adaptation of this circuit in an ultra-short-wave receiver is the RCA Victor receiver especially designed for police and other short-distance communication services which utilizes a circuit of the conventional type as shown schematically in Fig. 299.

This receiver employs three type 56 tubes, one as an oscillating detector, one as a quenching oscillator, and one as an audio oscillator.

A cathode short-circuiting switch (not shown) is in the detector circuit which, when closed, causes the circuit to function as a superregenerative receiver and which, when open, biases the detector enough to stop oscillation. Under these conditions the circuit operates as a biased detector and an audio amplifier for monitoring purposes.

High-impedance (20,000 ohms) headphones are used for both receiving and monitoring purposes and the output circuit from the last audio tube is designed accordingly.

IV. FREQUENCY MEASUREMENTS

290. Heterodyne Frequency Meter. *a. Description.*—Figure 300 illustrates the wiring diagram of type CGR-41 frequency meter developed for the U. S. Coast Guard service.

This instrument is of the heterodyne type and is used for checking the calibration of radio transmitting and receiving equipment on vessels and shore stations. The instrument must first be checked against a crystal-oscillator standard or some other frequency standard. This standard will hereafter be referred to as the *calibrator*.

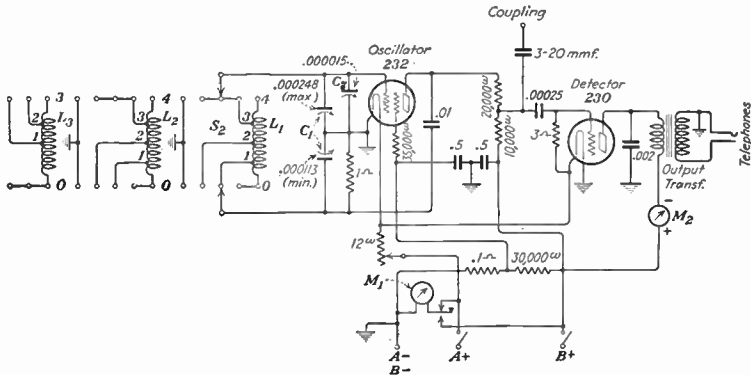


FIG. 300.—Heterodyne frequency meter. (Courtesy of U. S. Coast Guard.)

The frequency range of this instrument is from 95 to 550 kc./s. which may be obtained in eight steps, and an approximate range of 2,200 to 4,700 kc./s. in three steps.

The unit employs a screen-grid tube of the -32 variety as an oscillator and a triode detector of the -30 variety. A specially designed voltage-compensating circuit is provided in order that any variations in supply voltage may not appreciably affect the frequency. An actual variation of the supply voltage to + or -9 volts results in a frequency change of less than 0.01 per cent. Similarly, with a change of filament voltage of approximately 0.1 volt the percentage of frequency deviation will not be more than 0.01 per cent. If, therefore, a careful check of the filament and plate voltages is maintained, an extremely accurate check of the frequency may be had at all times.

The detector tube is connected to the oscillator by fairly loose resistance coupling and is also connected to the COUPLING binding post on the panel. Connection of any apparatus to the COUPLING post results in negligible frequency change for the oscillator, at any frequency within its range. The plate-current meter M_2 in the detector circuit indicates by its decreased reading whether or not the oscillator is functioning

properly. The oscillator may be stopped by turning the coil-selector switch to either of the two blank positions. If the detector-plate-current-meter reading is then noted, the change when the coil-selector switch is turned to any desired range is easily observed. A change of several milliamperes should be obtained on all points in the low-frequency range; a change of from 0.25 to 1.0 ma. should be obtained on all points in the high-frequency range. When a signal introduced at the coupling post is of sufficient intensity, visual beats are indicated by the detector plate-current meter, if the frequency of the heterodyne is within a few cycles of zero beat with the introduced frequency.

b. Theory of Operation.—The apparatus combines two essential elements, a stable oscillator of adjustable frequency and a detector, in a form convenient for use in frequency measurements. The oscillator is especially arranged for permanence (by rugged construction) and freedom from frequency shifts due to changes in supply voltages (by voltage compensation, described later). The detector is used to produce audible beats between the frequencies of the oscillator and some other oscillator (such as a transmitter or a separate crystal-oscillator calibrator).

The heterodyne oscillator circuit is a modified form of Colpitts oscillator, which is inherently one of the most stable types. The construction of the oscillating circuit is made as rugged as the requirements of size and weight will permit. Voltage compensation is obtained by properly proportioning the plate and screen-grid voltages of the oscillator tube, by means of a voltage divider, so that variations in total plate-supply voltages produce very small variations in frequency.

If two alternating voltages, of different frequencies, are impressed on a non-linear grid circuit (that is, a circuit which does not follow Ohm's law) the average grid current will undergo variations at a rate equal to the difference of the two frequencies. If the two frequencies are radio frequencies (above audibility) and the difference between them is made small enough to result in an audio-frequency beat current (within audibility), then the variations in grid bias take place at an audible rate. These variations are amplified by the tube, and, if telephones are connected in the plate circuit, the plate-current variations will produce a tone in the telephones. If the difference between the two frequencies is made very small (only a very few cycles), then the telephones will no longer respond to the difference, or beat, frequency. A meter placed in the plate circuit will indicate by pulsations these very low-frequency current variations, or beats, when necessary.

The compensating condenser, located near the coupling binding post, is provided for adjustment of the initial capacity of the tuned circuit. The condenser is provided with a screwdriver adjustment, access being obtained by means of a removable spring cover. If the heterodyne oscillator has been calibrated and a change in initial circuit capacity takes place, the calibration will of course be in error. Such changes may be in tubes, or due to slight displacement of the parts of the instrument relative to each other or relative to the case. Adjustment of the compensating condenser so that the reading of the instrument agrees with the original calibration, will bring all other points into agreement, provided the change in the circuit under measurement was a change in initial capacity. The compensator will not realign the instrument against changes in tube characteristics or changes in coil inductances, though adjustment of the compensator will aid materially in reducing shifts due to other causes than changes in the initial circuit capacity.

c. Practical Operation.—Shortly before the instrument is to be used for adjusting a receiver or a transmitter to a desired frequency, the heterodyne frequency meter

must first be checked against a crystal-oscillator calibrator. In checking, it is possible to listen for beats with the telephones connected to either the crystal-oscillator [calibrator] or the heterodyne frequency meter. The coupling between the calibrator and the frequency meter is obtained by connecting the COUPLING posts on both instruments together. If the coupling is too great, the wires may be left disconnected but close to each other, thereby providing slight electrostatic coupling.

If the crystal calibrator provides a crystal harmonic at a desired calibrating frequency, identify the harmonic by referring to the crystal-oscillator calibration. Then set the crystal-oscillator [calibrator] to zero beat with this harmonic. Turn off the crystal-oscillator [calibrator], and start the heterodyne frequency meter by closing the filament switch.

Select the proper coil of the heterodyne, as determined by reference to the range table on the panel (that is, the coil having a range in which are located the calibration points desired). Turn on the calibrator and adjust the variable condenser C_1 of the heterodyne meter until zero beat with the crystal-oscillator calibrator frequency is obtained. If the width of the zero-beat setting, as determined by headphones, covers several divisions on the condenser dial, a more accurate setting may be obtained by noting the pulsations of the detector plate-current-meter pointer, either side of zero-beat setting. If the signal is too weak to operate the meter needle in this manner, coupling between oscillators should be increased.

The frequency-meter oscillator may now be turned off, and the heterodyne frequency meter slightly readjusted to give zero-beat setting directly against the [calibrator] harmonic.

The frequency, coil, and condenser settings, and the temperature should be entered on the removable calibration card on the panel.

If the crystal calibrator does not provide a crystal harmonic at the desired calibrating frequency, first set the calibrator frequency to the desired frequency by interpolation between harmonics of the crystal, which occur near the desired calibrating frequency. Turn off the crystal oscillator.

Next, start the heterodyne frequency meter, check voltages, and note temperature. Select the proper coil on the CGR-41 heterodyne, as determined by reference to the range table on the panel. Then adjust the variable condenser, C_1 , until zero beat with the calibrator frequency is obtained.

The frequency, coil, and condenser settings, and the temperature should be entered on the removable calibration card on the panel of the frequency meter.

To Set a Receiver to a Desired Frequency.—Since the radiation from modern oscillating receivers is small, it is best to listen on the receiver. Start up the portable heterodyne frequency meter and adjust it to the checked frequency, making correction in the dial reading for any change in temperature of the instrument. Adjust the receiver to zero beat with the frequency of the frequency meter. Because of the comparatively great sensitivity of a receiver, very loose coupling between the receiver and the heterodyne frequency meter may generally be used. A wire connected to the heterodyne frequency meter COUPLING post and brought near the receiver antenna terminal is generally sufficient. If necessary, the wire (insulated) may be wrapped a few times around the receiver antenna lead to provide larger capacity coupling.

To Adjust a Transmitter to a Desired Frequency.—The headphones must be used with the heterodyne frequency meter for this operation. Start up the frequency meter as previously described, and adjust its setting to the desired checked frequency, making correction in this setting for any change in temperature of the instrument. Adjust the transmitter master-oscillator and antenna circuits, to zero beat with the

frequency-meter frequency. (This presumes that the master oscillator and antenna circuit are to be operated on the same frequency without doubling.)

A short piece of insulated wire, connected to the COUPLING post of the frequency meter should suffice for a pick-up circuit. In a well-shielded master oscillator, this coupling wire may have to be brought close to the master oscillator circuits, and, of course, all other circuits should be placed out of resonance while calibrating the master oscillator. However, slight readjustments to the master oscillator may be required after all circuits are resonated due to coupling effects between the transmitter circuits. Checking the transmitter output frequency against the calibrator should be the final operation, and the output should be at zero beat on the heterodyne frequency meter when all circuits are finally in resonance. This final measurement of the antenna circuit should be made with very loose coupling to the heterodyne frequency meter, which will require no coupling wire, and with the frequency meter placed at a remote location in the radio room.

CHAPTER XII

RADIO AIDS TO NAVIGATION

Although directional transmission and reception characteristics were noticed by Hertz in his early experiments, no practical use was made of the phenomena until the World War when the radio direction finder was developed for use in locating enemy radio stations. Since that time it has found extensive use, particularly in marine navigation.

I. PRINCIPLES OF RADIO DIRECTION FINDERS

291. Fundamental Principles of the Loop Antenna.—The operation of the radio direction finder depends on the directional characteristics of the loop antenna. These directional properties of a loop may best be understood by studying a simple experiment. Imagine a loop antenna (Fig. 301) consisting of several turns of wire wound around a rectangular form, say 5 ft. across, with its terminals connected to a sensitive thermogalvanometer. Assume further that the loop is mechanically arranged so that it can be rotated on a vertical axis XY represented by the dotted line.

If the loop is now placed in an open space away from any surrounding objects, and a reading taken of the current induced in the loop by a distant transmitter, it will be found that (a) the current is minimum when the plane of the loop is at right angles to the line of direction from which the signals are emanating, and (b) the current is maximum when the plane of the loop is parallel to the line of direction taken by the signals.

The effect on the current flow in the loop caused by orienting the loop to the signal may be graphically expressed as shown in Fig. 302. This is called a *figure-of-eight* or *bilateral characteristic*. The lines radiating from the center show the relative current values. These values are expressed as units by the numbers at the end of each line. When the ends of these lines are connected a figure-of-eight curve results.

A study of this curve (Fig. 302) will show that, when the waves are approaching, or traveling along, a line drawn through points I and III, at right angles to the stationary plane of the loop, the current in the

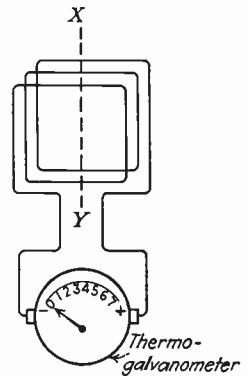


FIG. 301.—Simple loop circuit.

loop is at a minimum value. As the line of travel followed by the waves changes clockwise from the I-III direction, the current in the loop increases. At 45 degrees the current value has reached almost 12 units. At 90 degrees, the point of maximum current flow in the loop, a value of 16 units is reached. The waves are now traveling along the II-IV line which is parallel to the plane of the loop as shown in the figure. Theoretically this is the position at which a maximum current should be induced in the loop, as will be explained later.

As the line along which the waves travel is oriented farther in a clockwise direction the current commences to decrease until a new minimum is found when the line of wave travel has changed its position, relative to the loop, by 180 degrees. The line of wave travel is now from

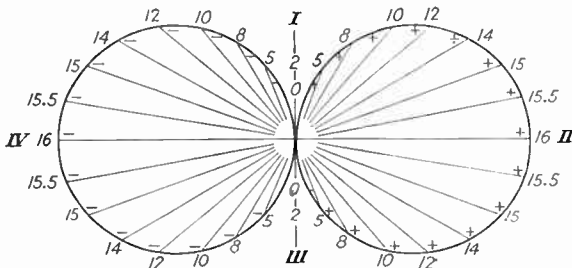


FIG. 302.—Bilateral characteristic of loop.

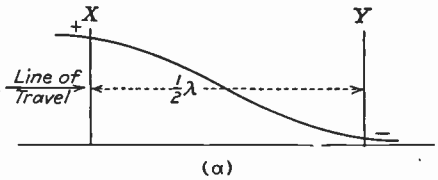
points III to I, the reverse of I to III. As orientation continues, the current increases to a new maximum at 270 degrees, and then decreases to minimum again when the rotation has been completed.

The points of minimum signal are critical, far more so than the points of maximum signal, and these points may therefore be used to determine the *line* of direction taken by the traveling radio waves.

One way of accounting for the directional properties of the loop is as follows: When the wave is approaching along a line perpendicular, or at right angles, to the plane of the coil, the wave front is parallel with the plane of the coil. Therefore, the wave strikes both sides of the loop at the same instant (the horizontal wires contribute nothing to the directional properties of the loop) and induces e.m.f.s. of identical phase in both sides of the loop which cancel out, nullifying any tendency for a current to flow. If, on the other hand, the wave is approaching along a line of travel parallel with the loop, the wave front strikes one side of the loop a fraction of a second sooner than it strikes the other side. Although this fraction of time may be only a part of a millionth of a second, the speed-of-current effect is so great that the current in the side of the loop nearest the source starts to flow before the current in the side farthest from the source has time to build up to sufficient proportions to nullify it.

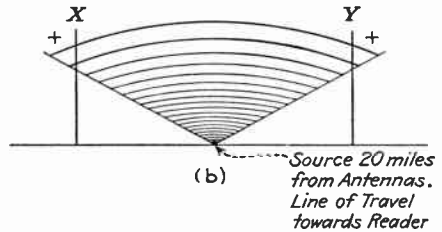
This action may be better understood if we imagine a slow-motion picture of the process. Suppose two vertical wires are set up in a desert spaced 300 meters apart. This distance is chosen because at a speed of 300,000,000 meters per second it will take a 600-meter radio wave 0.000001 sec. to strike the second wire after it has struck the first if the line of travel is parallel with an imaginary line stretched between the two wires.

If the line of travel is across the page and the source is to the left of the diagram, as in Fig. 303(a), the e.m.fs. induced in the wires by the 600-meter wave will be 180 degrees out of phase, that is, when the polarity is maximum positive at the top of wire X, it is maximum negative at the bottom of wire Y. If now the tops of the two antennas are connected by a wire, and connecting wires run from the bottoms of the two antennas X and Y to a sensitive thermogalvanometer, a current will flow in the thus formed loop and will be registered on the meter. The horizontal wires at the top and bottom of the system contribute nothing to the e.m.f. in the circuit. It is found that a maximum current is induced in X and Y when the waves are approaching along a line parallel with a line drawn between the two wires.



(a)

If X and Y are closer together, the phase angle between the induced currents is relatively less; otherwise, the effect described is the same regardless of the length or spacing of the wires.



(b)

FIG. 303.—Effect of approaching wave on two vertical antennas.

If the same circuit set up is allowed to remain, and the source of signals shifted to a point at right angles, or perpendicular to a line drawn between the two wires X and Y, as shown in Fig. 303(b), the crest of the wave will strike the two antennas at the same time thereby inducing identical in-phase currents in both which cancel out by virtue of their bucking effect. No current would be registered on the meter under these conditions.

In the practical example just cited it was assumed that the loop circuit was free of surrounding objects which would distort the oncoming wave. Under practical working conditions these ideal conditions are not usually found, and it is rarely possible to get a sharp minimum signal when the loop is at right angles to the source unless special precautions are taken in the design and installation of the system.

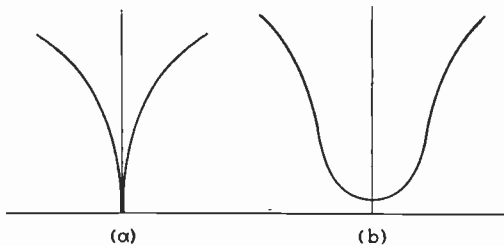


FIG. 304.—Signal characteristic: (a) ideal condition of sharp minimum; (b) condition with residual signal present, broad minimum.

When the point of minimum signal is not critical, there is said to be *residual signal* present. This may be shown graphically as in Fig. 304, where (a) is the ideal condition of sharp minimum, and (b) is the condition with residual signal present. This residual signal may be caused by conditions within the direction-finding equipment itself which cannot be remedied except by changing electrical conditions in the direction-finder

circuit and equipment, or it may be caused by conditions extraneous to the direction-finder circuit in which situation no amount of rearranging the direction-finder circuit will effect a remedy.

Let us first consider inherent circuit deficiencies which might cause residual signal. The principal trouble is that caused by *antenna effect* in the loop. Antenna effect is the result of the flow of dielectric currents from various parts of the loop circuit (and from receiving circuits coupled to the loop) to the ground. Some of these paths may not lead directly to the ground but form short cuts to the ground with the same effect. Some of these dielectric currents, as they might flow from an unshielded loop, are shown in Fig. 305 as dotted lines.

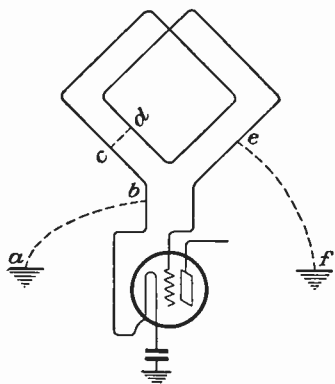


FIG. 305.—Dielectric currents (dotted lines) causing antenna effect.

Line *ab* represents the leakage from connecting leads to the ground; line *cd*, the short-cut leakage between the turns in the loop; and line *ef*, the leakage from the loop itself to surrounding objects which are grounded, or directly to the ground itself. There are probably many other paths, but these are the principal ones.

The e.m.fs. produced by the antenna effect are generally 90 degrees out of phase with the e.m.fs. produced by the coil and for this reason can never completely balance out the coil e.m.fs. Also the e.m.fs. produced by antenna effect do not vary with orientation of the coil, or, in other words,

no matter how the coil is turned, the antenna effect for a given coil and circuit is the same.

The other principal cause of residual signal or broad minimum is the e.m.fs. induced into the direction-finder circuit by spurious fields produced by near-by metallic objects which form closed-loop circuits. The energy is transferred from these outside circuits by induction, and not by radiation, which, from these circuits, is too feeble to be noticeable. Because the energy is transferred by induction, the potential of the e.m.fs. induced depends upon the orientation of the coil. The effect then varies with coil position and usually results in a point of minimum signal which is different from that which would result if the coil was not affected by these spurious fields.

Another source of error in finding the position of minimum signal is introduced by a so-called *night effect*. This effect is noticeable during the period continuing from $\frac{1}{2}$ hr. before sunset to $\frac{1}{2}$ hr. after sunrise and is probably caused by the erratic nature of the wave propagation during the night period. Experience has brought out the fact that this night effect is most pronounced shortly before and after sunrise or sunset.

Errors in the position of minimum signal are manifest by a rapid swinging of the minimum, which gives the effect that the source is changing its location while the observations are being made. Freak minima sometimes exist as part of the phenomena, in which the null (point of minimum signal) is noted as merging into several distinct nulls, and again it may become entirely extinct. Considerable residual signal usually accompanies night effect, and bearings taken under these conditions are to be considered as being only approximately correct. At distances less than 100 miles, the error introduced is negligible. It has also been noticed that at times a variable error due to night effect might be reported by a shore direction-finder station while no error is noticed by ships in the immediate vicinity equipped with modern direction finders.

The cause of the so-called *night effect* is an erratic variation in the relative strength of the sky wave at night compared to the ground wave. This results in strong sky waves which are horizontally polarized, that is, their electric field and wave front are parallel, or tend to take this position, to the earth's surface. Therefore, when they strike the horizontal wires of the loop, these waves set up a current flow in these wires. These currents tend to nullify the currents flowing in the vertical wires of the loop, which are due to the effect of vertically polarized ground waves and to space waves traveling along the surface of the earth.

As the loop of a radio direction finder is turned on a vertical axis, the relative position of the horizontal loop wires and of the down-coming horizontally polarized sky waves is unaffected by rotation of the loop, and, therefore, the night effect is present regardless of the position of the loop.

As a matter of fact, the horizontal polarization of radio waves takes place during daylight hours as well as at night, but the effect during the daytime is negligible owing to the relative weakness of the sky wave compared with the ground wave on frequencies generally used for direction-finding purposes.

The only remedy for night effect over distances in excess of 100 miles on optimum frequencies is to concentrate the radiation from the transmitter into ground and vertically polarized space waves as is done by the transmission-line antenna used for aeronautical radio-range beacons.

Just as light waves are bent from their normal path of travel when they pass from a medium of one density to a medium of another density, so are radio waves bent when they leave the medium of travel over water and enter an adjoining medium which is over land. For this reason a considerable error might be introduced into any bearing taken by a radio direction finder on a signal that has traveled for an appreciable distance along a coast line owing to the difference in the media of travel over water and land. Any line of direction indicated by a radio direction finder under these conditions should be considered inaccurate. For the same

reason a bearing taken when the radio direction finder is separated from the source by intervening land should be regarded as more or less approximate.

Any distortion of the radio wave before it strikes the loop, be it caused by induction from near-by objects, night effect, or anything else, is referred to as *deviation*. From the discussion thus far it is seen then that residual signal is generally caused by either antenna effect or deviation.

The error caused by residual signal may be plotted graphically into a curve which is known as a calibration chart. Such a chart, which accompanies every radio direction-finder installation, is shown in Fig. 306.

To summarize the principal difficulties experienced when the loop is put to practical use as a direction finder, it may be stated that *antenna*

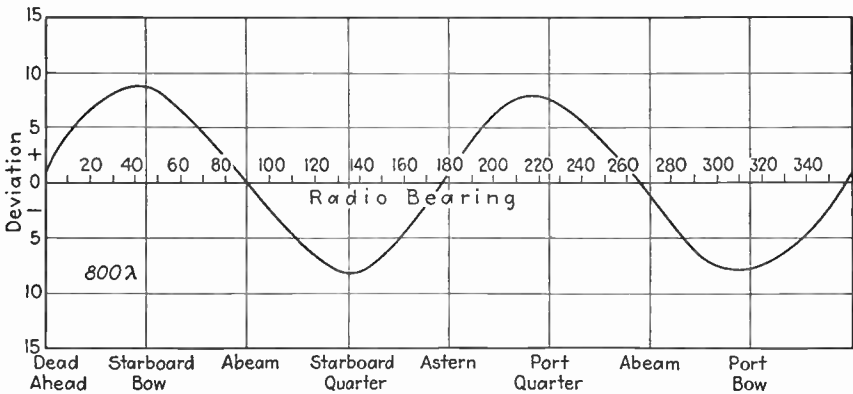


Fig. 306.—Calibration chart for radio direction finder.

effect and *deviation* are the chief problems to be solved by the designing and installation engineer.

292. Minimizing the Error. a. Reducing Antenna Effect.—The absence of a critical point of minimum signal is mainly due to residual signal which, as has before been pointed out, is that signal which is sometimes present in the arc that should be void of all signal. From a practical standpoint this means that the operator cannot get a minimum signal, no matter how he turns the loop. This problem presented a great difficulty in the early development of the radio direction finder for marine applications, when large unshielded loops were used.

Upon investigation it was found that residual signal was caused largely by two things, namely, (a) direct pick-up due to induction from near-by metallic circuits, and (b) antenna effect due to lack of symmetry on the part of the direction-finder circuit, and dielectric currents.

Both of these undesirable characteristics were minimized by shielding the loop, loop leads, receiver, and all accessories. The continuity of the loop shield was broken at the apex of the loop by insulating fittings.

To improve the electrical symmetry of the loop circuit, push-pull radio-frequency amplification was employed. These were obvious solutions which soon occurred to the designing engineers, but they did not completely solve the problem. Various methods of balancing out the undesired residual signal by the use of balancing circuits were then tried.

The object of the *balancing circuit* is to introduce currents of a definite phase value into the loop circuit, thus balancing out the antenna effect and consequently a large part of the residual signal. The final form of this balancing circuit is shown in Fig. 307. The condenser C_1 tunes the loop to the frequency of the incoming signal thus making the circuit more responsive (increases pick-up) than if the loop circuit were untuned. The

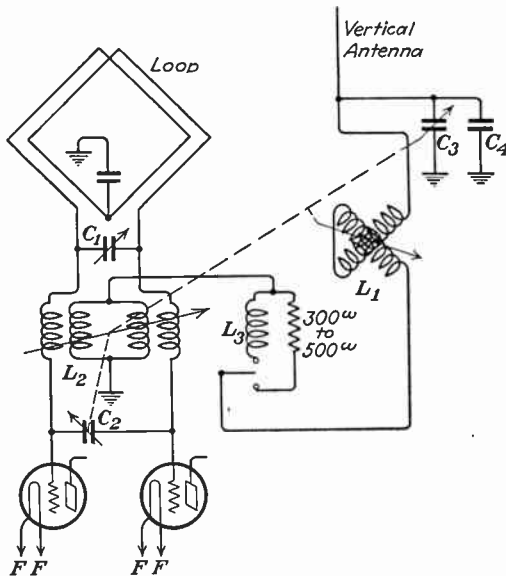


FIG. 307.—Fundamental balancing circuit employed to reduce antenna effect.

antenna tuning condenser C_3 does the same thing for the antenna circuit as shown. The antenna circuit is also tuned by the variometer L_1 .

The size of the balancer antenna should be such that it will pick up sufficient e.m.f., which, when used with proper L , C , and R values, will balance out the antenna effect. On some commercial installations the balancer antenna is a vertical wire about 50 ft. long, raised up about 35 ft. and guyed off to one side to give it rigidity. On other installations smaller antennas (15 to 25 ft.) have been found sufficient. The exact size is usually not critical and can be found by a little experimentation.

Condensers C_2 , C_3 , and inductances L_1 and L_2 operate in tandem together with the radio-frequency tuning condenser and radio-frequency oscillator condenser (if a superheterodyne receiver is used). This arrangement provides proper tuning values for all the circuits over the entire wave-

length range of the equipment. The balancing e.m.f. is transferred to the loop from the balancer antenna by the inductive coupler L_2 . The two windings on the rotor of L_2 are wound in opposite directions, thus producing opposing fields and eliminating any tendency on the part of L_2 to tune the loop circuit.

On large vessels where the loop can be placed relatively free from surrounding objects the means for eliminating antenna effect just described are generally sufficient. On the other hand, on smaller vessels and on some large ones, lack of space makes it necessary to place the radio direction-finder loop in the midst of metallic circuits such as stays, smokestacks, iron railings, and so forth. Under these conditions additional precautions and methods of eliminating deviation must be taken.

b. Reducing Deviation.—On small vessels where a clear open space cannot be found for the loop, serious deviation may result owing to

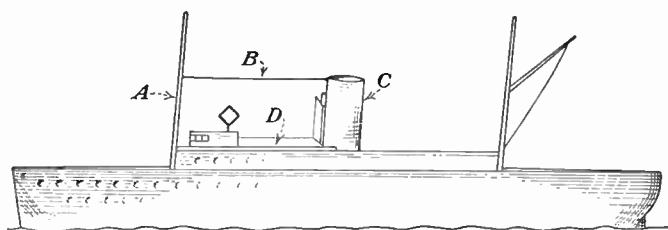


Fig. 308.—Use of compensating loop (A, B, C, D) to reduce deviation.

spurious fields set up by surrounding metallic objects. If the deviation thus set up is more than 20 degrees, it is considered excessive and steps must be taken to reduce it to normal. This may be done by the use of *compensating loops* which are formed by connecting together the various metallic circuits so that they form closed loops as shown in Fig. 308. Side A of the loop is formed by the metal mast; side B, by a wire connecting the mast and smokestack; side C, by the smokestack; and side D, by the iron railing which is electrically connected to the mast and stack.

The electrical dimensions of the compensating loop are important. Care must be taken not to make the compensating loop resonant with the direction-finder circuit. If this happens, the deviation will be enormously increased. On the other hand, if the compensating loop is not large enough, an overcompensated condition will result which will spread the deviation over a wider arc than when the proper compensating loop is used. In addition to altering the electrical dimensions of the compensating loop by varying its mechanical size, the electrical characteristics of this loop may be varied by the insertion of lumped resistance, inductance, or capacity in series with it. For example, a resistance, coil, or condenser may be inserted in series with the wire B of Fig. 308.

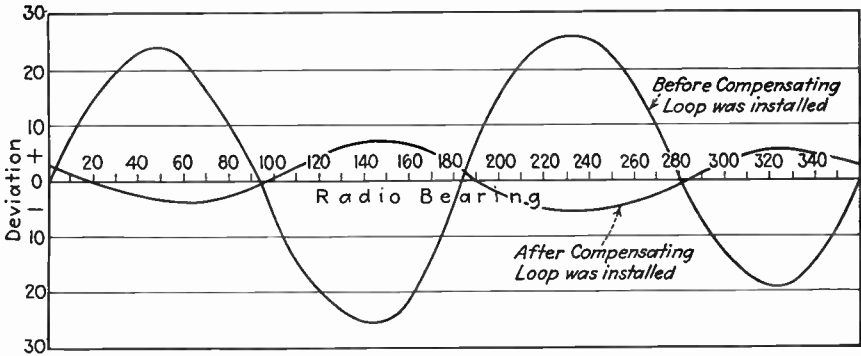


FIG. 309.—Error curve before and after installation of compensating loop.

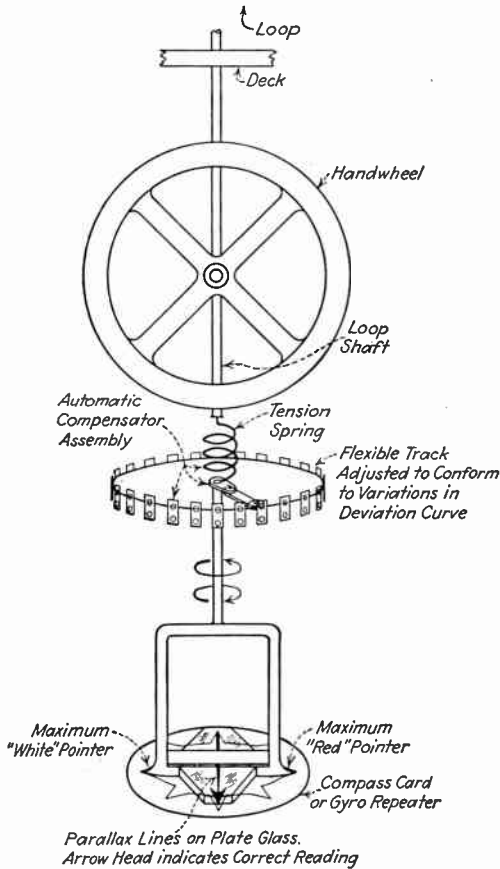


FIG. 310.—Mechanical arrangement of automatic compensator to loop shaft.

Typical effects before and after the installation of a compensating loop are shown by the two curves in Fig. 309.

If only normal deviation is encountered, this is compensated for by a mechanical compensator which automatically causes the indicator to lag or lead the plane of the direction-finder loop by the necessary amount. The mechanical idea of this compensator is shown in Fig. 310.

After the radio direction finder has been calibrated and placed in operation, care must be taken to see that no large masses of metal in the vicinity are shifted in position, as this would probably affect the calibration. No local radio antennas should be moved (watch broadcast-receiver aerials too), and all antennas should be open-circuited (not grounded) while the radio direction finder is in use.

On shipboard installations, the average deviation, as shown in Fig. 309, is about as follows, starting at the bow and rotating clockwise.

Degrees (approximate)	Deviation
45.....	Maximum (positive)
90.....	Minimum
135.....	Maximum (negative)
180.....	Minimum
225.....	Maximum (positive)
270.....	Minimum
315.....	Maximum (negative)
360.....	Minimum

293. Effect of Frequency on Deviation.—Various experimenters have made different reports on the effect of frequency on deviation. It is generally conceded by authorities in the United States that deviation is affected by frequency. Since its invention as a practical navigation aid the marine-type radio direction finder has given very reliable service up to distances approximating 100 miles. At distances up to 50 miles an average accuracy to within 2 degrees has been noted for bearings taken in the bow and stern quadrants, and an accuracy of from 2 to 4 degrees in the beam quadrants when the signals are reasonably strong (*QSA* 4 to 5, see end of Appendix B). This statement applies particularly to the 800-meter radio-compass band but is *probably* true for a wave-length range of from 200 to 1,000 meters.

For many years it was believed that the deviation on high-frequency emissions below 100 meters was too great to be compensated for owing to the vertical polarization of these waves which caused the electric component of the wave to lie in the vertical plane of propagation. Thus the majority of the energy reaching the direction-finder loop would be reflected from the Heaviside layer, and any directional effect of the loop would be extremely unstable. This has been found to be not wholly

true, and later experiments (beginning in 1926) have demonstrated that, although the deviation is greater at frequencies above 100 meters, the radio direction finder may be reliably used up to frequencies as high as 7,700 kilocycles (39 meters) and even up to 12,000 kilocycles (25 meters) if a good strong ground wave can be received. Directional qualities at the higher frequencies are sufficiently stable to enable fairly accurate readings in the quadrantal sectors with maximum deviation occurring halfway between these points.

294. Determining Sense of Direction.—The inherent characteristics of the loop are bilateral in nature, and, therefore, only the *line* of direction taken by an observed signal can be determined by the loop when used alone. It is important, especially on shipboard, to be able to tell from which quadrant a signal is coming. This is accomplished by mixing energy from the balancer antenna (vertical wire) with that taken up by the loop in such phase relation that, when both e.m.f.s. are added vectorially, the e.m.f. passed on to the receiver is greater when the loop points in one direction than when it points in the opposite. This is done by connecting an impedance circuit (consisting of L_3 and a 300- to 500-ohm resistance, Fig. 307) in series with the antenna by means of a control switch. To understand this it may be imagined that an indicating arrow is fastened to the top of the loop, parallel to the plane of the loop. Then, when the loop and vertical-antenna energies are mixed, the resultant signal is stronger when the arrow points, let us say, toward the source than when it points away in the opposite direction.

The e.m.f. conditions present in the loop circuit when the vertical antenna is coupled to it are called the *unidirectional* or *unilateral characteristic* of the loop which is graphically expressed by a cardioid curve as shown in Fig. 311.

The minus signs shown around the perimeter of the left-hand circle of the figure-of-eight characteristic of the loop, as shown in Figs. 302 and 311, indicate that the phase conditions reverse when the source of the signal with relation to the loop is reversed. It is to be noted that the positive signs around the cardioid characteristic in Fig. 311 are the vectorial result of adding the positive or negative values of the figure-of-eight characteristic to the positive values of the circular characteristic curve which represents the energy from the vertical antenna. This antenna has non-directional characteristics, and, therefore, the phase relationships throughout the entire 360 degrees may be represented as positive. *Under proper conditions of operation* the relatively positive energy in the antenna is always at least equal to any relatively negative energy which may be generated by the loop, and the cardioid characteristic is always positive in nature. If the pick-up from the vertical antenna is not sufficient, a non-perfect cardioid characteristic will result, as shown in Fig. 312.

295. The Bellini-Tosi System.—This system of radio direction finding was named after its inventors, two Italian investigators. The system utilizes two stationary loop antennas which are erected at right angles and connected to an instrument box which contains a so-called *radio goniometer*. The arrangement is illustrated in Fig. 313. The familiar directional properties of a loop antenna are utilized here, but this system differs from the conventional radio direction finder using the rotating small loop in that in this system the antenna is stationary, angular displacement variations being taken care of by the radio goniometer, which is the receiver in this system.

The radio goniometer consists of two sets of stationary coils *AA* and *BB* (Fig. 313) and a movable exploring coil *C* mounted in a suitable cabinet together with the necessary detector and amplifiers and connected to the loops as shown in the diagram. This box may be placed in the chart room or pilot house so that the navigator may operate the instrument himself.

Let us assume for purposes of explanation that the coil aerial *a* is parallel to the keel of the ship or fore and aft, and that the coil aerial *b* is athwart-ship or at right angles to *a*.

In accordance with the well-known directional characteristics of loop aerials, a maximum signal is induced in a loop when the loop lies in a line of direction parallel to the direction of propagation of the oncoming wave. Also, a minimum signal is induced in a loop when the loop lies in a line of direction at right angles to the line of direction of propagation of the oncoming wave.

A study of the diagram will show that a current flowing in loop *a* must also flow in coil *A*. Likewise a current flowing in loop *b* must flow in coil *B*. Therefore, the current flowing in the coils *A* and *B* is in reality the current induced in the loops *a* and *b*. If a strong signal is induced in *a*, a strong signal flows in *A*. If a strong signal is induced in *b*, then a strong signal flows in *B*. Therefore, the field set up in the

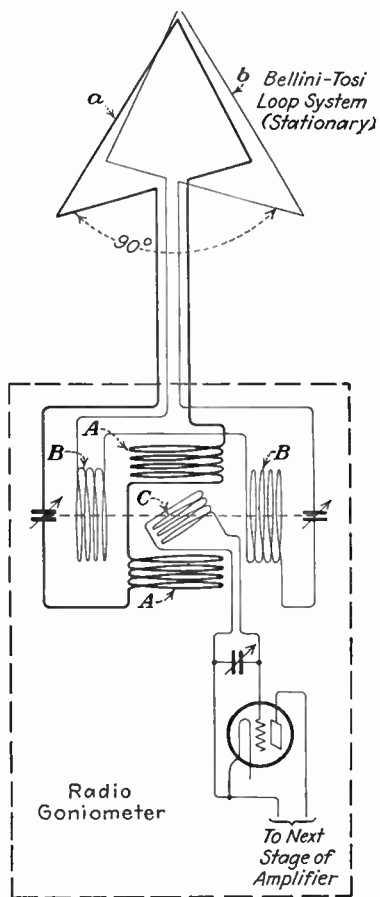


FIG. 313.—Radio goniometer utilizing Bellini-Tosi loop system.

space around the exploring coil *C* is similar to the field set up around the loop aerials *a* and *b*.

A maximum e.m.f. is induced in the exploring coil when that coil is at right angles to the magnetic field set up around it.

Now suppose a signal comes from the direction in which the ship is headed. A maximum signal is induced in *a* and flows in *A* also. If the coil *C* is oriented so that it is parallel to *A*, then a maximum e.m.f. will be induced in it and a maximum signal will be passed on to the receiver. If the coil *C* is now placed at right angles to *A* or parallel to *B*, a minimum signal will be passed on to the receiver.

The same action takes place if a signal comes from abeam. In this case a maximum current flows in *b* and *B* and a minimum current in *a*

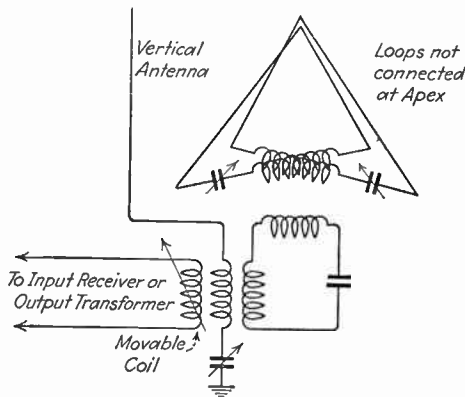


FIG. 314.—Scheme for connecting non-directional vertical antenna to the Bellini-Tosi loop system.

and *A*, and the receiver responds accordingly as the exploring coil is turned parallel to *B* for maximum response and at right angles to *B* for minimum response.

Now suppose a signal approaches from 45 degrees off the port bow. There are now equal e.m.f.s. induced in both loops *a* and *b*. The fields set up by coils *A* and *B* are of such phase relationship that a maximum e.m.f. is induced in coil *C* when it lies 45 degrees off the parallel position to both coils *A* and *B*.

A pointer is attached to coil *C* which moves over a pelorus scale which shows the bearing of the transmitting station, and its reciprocal, with respect to the ship. If the ship is kept on the course and another bearing taken later, these relative bearings, together with the distance traveled, may be computed to give the exact position of the observing station, as explained under Sec. 304, Fixing the Position.

A scheme for coupling a non-directional vertical antenna to a Bellini-Tosi loop system to get a unidirectional characteristic is shown in

Fig. 314. The theory underlying this action has been explained earlier in this chapter.

The chief disadvantage of the Bellini-Tosi loop system is the size of the loops which must be used in order to get sufficient pick-up to energize the receiver. When used for transmitting, the large loops used in this system are an advantage over a smaller loop because of the stronger field radiated by the larger loops. The Bellini-Tosi system has found some use on European ships for direction-finding purposes, but it has never been used to any extent for this purpose in the United States. A modified Bellini-Tosi loop transmitting-antenna system was used on American airways for radio-range transmitters but they have been replaced by transmission-line antennas which are free from night effects, and which will be described later.

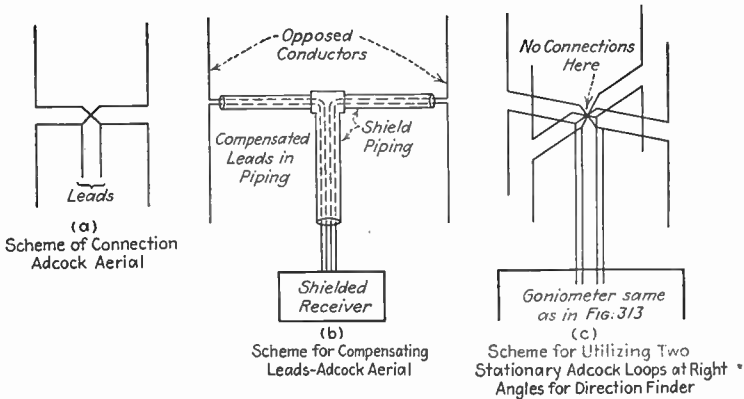


FIG. 315.—The Adcock aerial for directional reception.

296. The Adcock Aerial for Directional Reception.—This system utilizes a pair of spaced vertical conductors placed end to end as shown in Fig. 315(a) and (b) and mechanically arranged so that they may be rotated. The vertical conductors are the only effective conductors, as the horizontal sections are transposed so that compensation results and any energy picked up by them is neutralized. The same scheme is followed in bringing down the leads from the loops to the goniometer box.

The principal advantage claimed for this system is its freedom from night effects. The principal disadvantage is that loops of large dimensions are required if the pick-up is to be sufficient for practical work. If desired, the loops may be stationary, with two loops used at right angles to one another and connected to a goniometer as shown in (c).

The Adcock aerial system gives a symmetrical figure-of-eight characteristic, its theory of operation being similar to that of the closed loop described earlier in this chapter. This system has received considerable

attention in Great Britain but little has been done with it in the United States except that it formed the basis for the development of the transmission-line antenna system used with radio-range beacons.

II. THE ROTATING-LOOP TYPE RADIO DIRECTION FINDER

297. Applications of the Radio Direction Finder.—There are various ways of applying the radio direction finder to navigational purposes. The radio direction finder may be installed on a vessel or aircraft, and bearings taken on any source of signals. The advantage to the marine navigator of being able to do this himself in times of thick weather at sea is apparent. Likewise, when the instrument is installed in an aircraft, it enables the pilot to guide himself to a point where radio-range beacons (to be described later) are not available, or to determine the position of the aircraft.

To facilitate the taking of accurate marine bearings, there has been established by the U. S. Lighthouse Service almost a hundred radio beacons on most of the important lighthouses and lightships on both coasts of the United States and on the Great Lakes as shown in Figs. 316, 317, and 318.

Radio beacons are transmitters which radiate equally well in all directions, and, as their position is accurately given in the publication "Commercial and Government Radio Stations of the United States," any bearing taken on these beacons may be considered as being accurate bearings from a given point. By taking bearings from two such points the exact location of the observer may be ascertained by cross triangulation.

Regular radio-beacon stations emit a characteristic signal which can be readily distinguished by any one without a knowledge of the International Telegraph Code. They are so located that bearings taken on them by vessels will in most cases be entirely over water. Radio-beacon stations operate continuously in thick or foggy weather, and at other regular intervals as set forth in the above mentioned publication.

A system consisting of about 50 radio direction-finding stations is maintained by the U. S. Navy at the entrances to the principal harbors on both coasts. More than half of these will transmit as a radio beacon on 375 kilocycles (800 meters) upon request. These stations are also listed in the publication referred to before. Any ship fitted with a radio compass may, therefore, take a bearing on any of these stations.

It is also possible for a vessel fitted with a radio direction finder to take a bearing on any commercial or government shore station which bearing might be useful provided the exact geographical location of the transmitting aerial is known. This practice is not recommended unless it is impossible to take a bearing on a radio beacon.

The navigator aboard a ship fitted with a direction finder may take bearings on another ship, a possibility which is especially useful when proceeding to a ship in distress or to any ship with which it wishes to make contact.

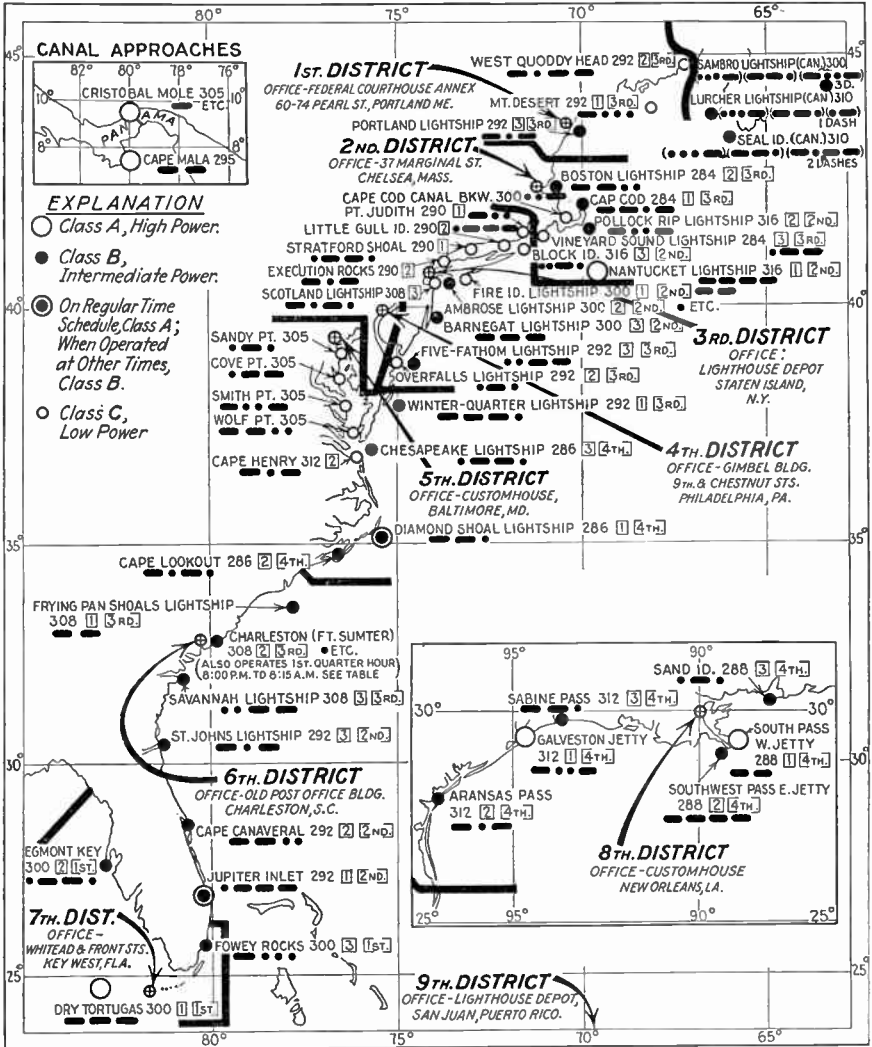


Fig. 316.—Atlantic Coast radio beacons. (Courtesy of U. S. Lighthouse Service.)

All of the applications thus far described would also apply to a seaplane, an amphibian plane, or a dirigible flying over navigable waters.

Itinerant airplane pilots may utilize existing broadcasting stations as radio beacons by the use of a specially developed unit which operates with a single small loop, no vertical antenna being necessary, and which

gives an automatic bilateral indication of the course, together with a directive sense, by a pointer instrument. This unit may be added to any radio receiver and operates equally well on modulated or unmodulated

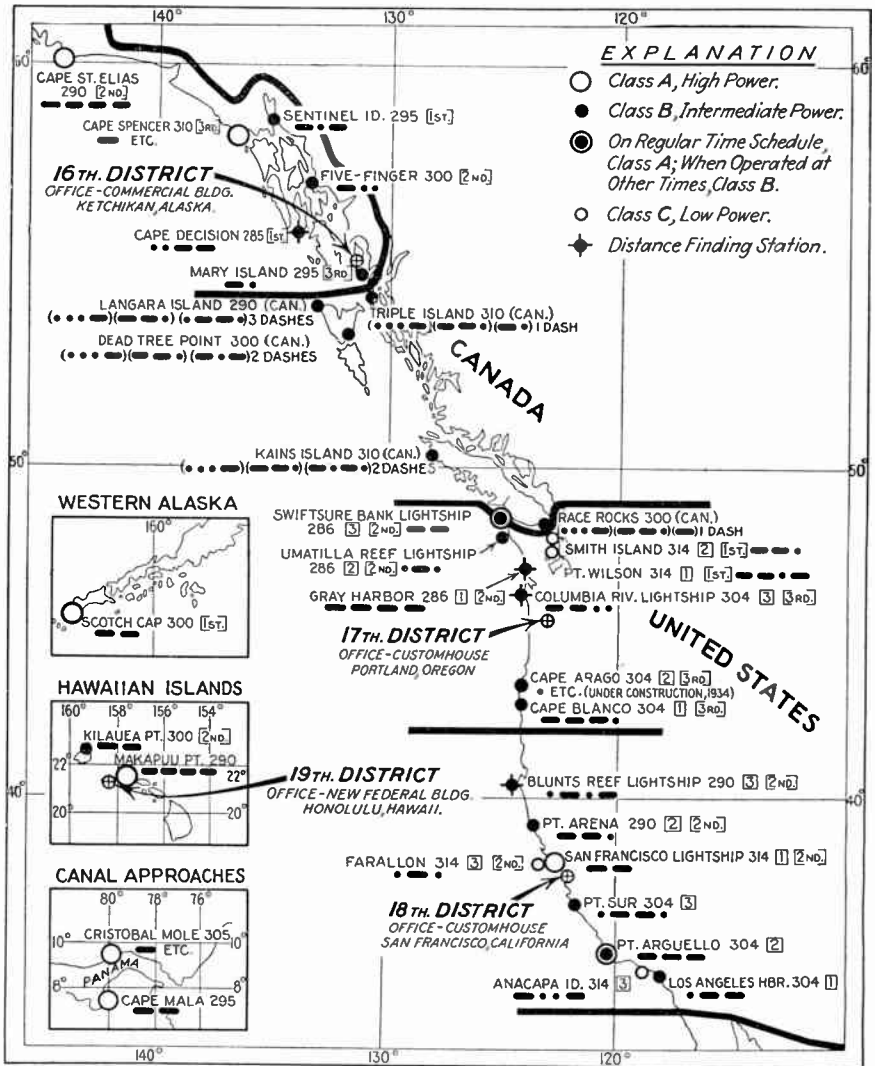


Fig. 317.—Pacific Coast radio beacons. (Courtesy of U. S. Lighthouse Service.)

signals. This direction finder, which is especially built for airplanes, will be described later in this chapter (see Sec. 311).

The marine-type direction finder is not automatic in operation and, therefore, differs in operation and use from the automatic device designed especially for the airplane. Direction finders to be used on transoceanic

aircraft, be they of the lighter-than-air or of the heavier-than-air types, will most likely be a combination of the marine-type direction finder and the airplane automatic type, and be semiautomatic in operation.

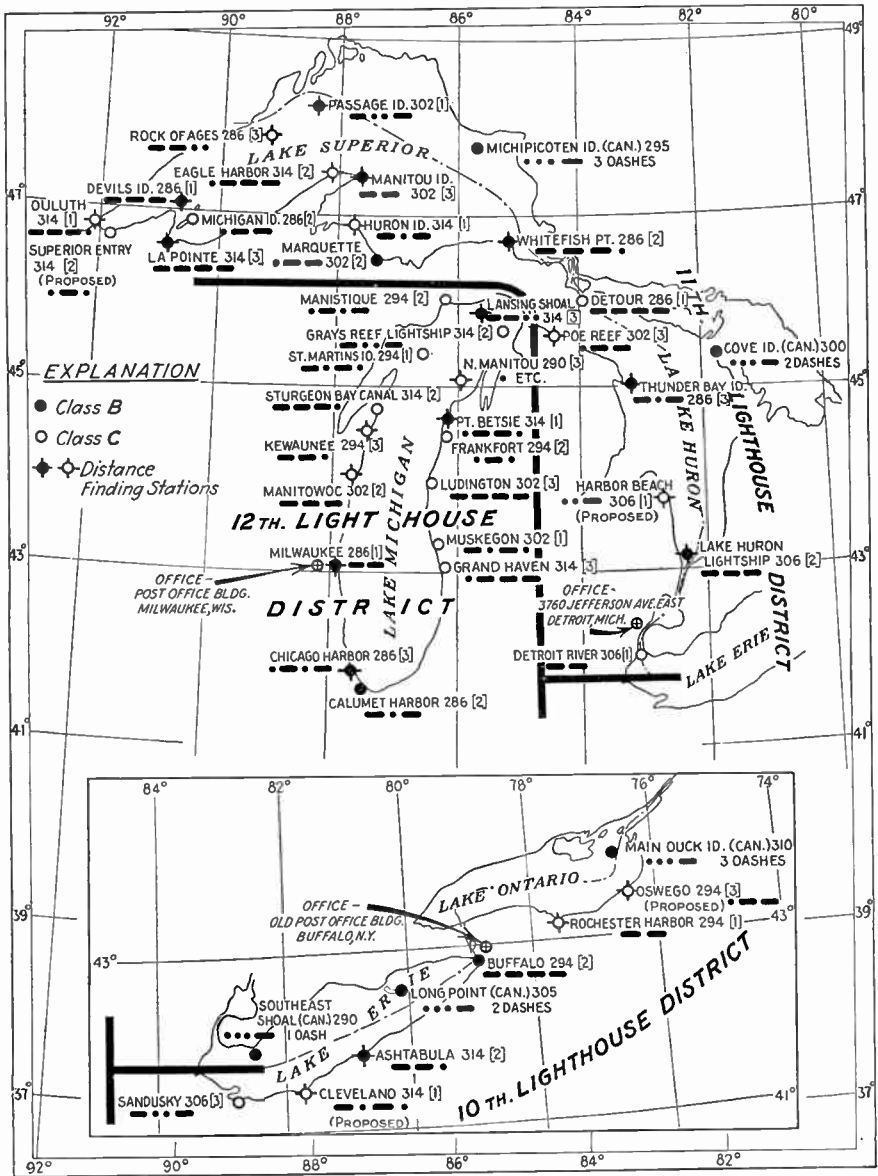


FIG. 318.—Great Lakes radio beacons. (Courtesy of U. S. Lighthouse Service.)

298. Elements of Direction-finder Apparatus. a. Description.—The radio direction-finder installation [Fig. 319(a)] consists of the loop, a

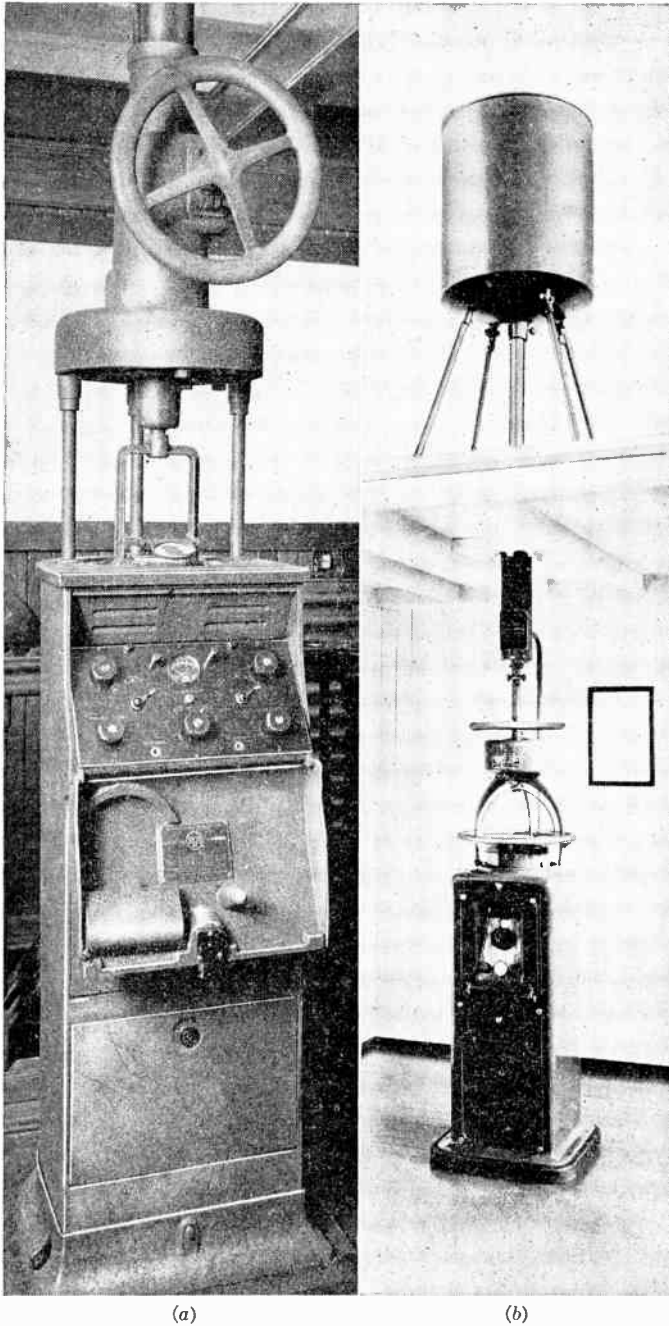


FIG. 319.—Marine-radio direction finders: (a) R.M.C.A. type; (b) Mackay type. [Photo (a) Courtesy of R.M.C.A.; photo (b) Courtesy of Mackay Radio and Telegraph Company.]

bearing indicator, a receiver, and a signal-intensity indicator which may be of the audio (headphones) or visual (meter or lamp) types. A mechanical drawing of the assembled apparatus is shown in Fig. 320.

The loop is mounted on a pedestal above the deck or radio direction-finder housing and is arranged so as to permit complete and continuous orientation about its vertical axis by means of a handwheel and a set of reducing gears. The loop is generally wound with a special radio-

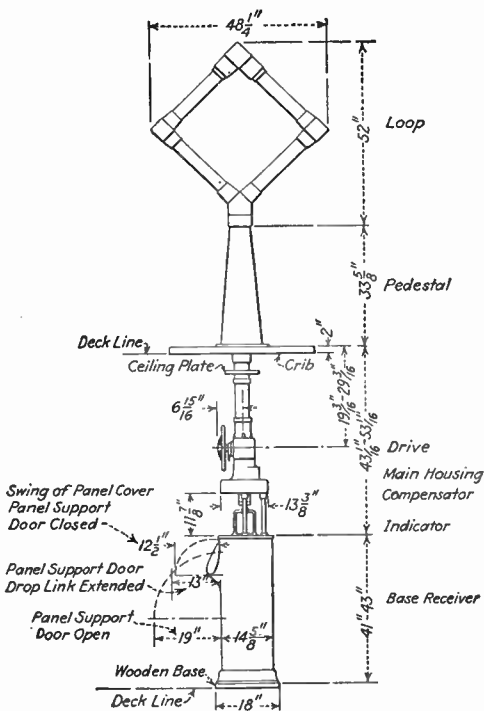


FIG. 320.—Mechanical arrangement of marine-radio direction finder, R.M.C.A. type.

frequency cable, the number of turns depending on the wave-length range to be covered and the size of condenser used with the loop. A loop 4 ft. on a side, wound with eight turns of No. 20 wire, or its equivalent in a stranded cable, when used with a 0.001-mf. variable condenser will tune to the 800-meter direction-finder wave.

b. The bearing indicator varies with the type of direction finder used. On direction finders designed for installation on large vessels engaged in deep-water navigation the indicator consists of a dumb-compass card mounted so that readings may be taken in degrees from the ship's head, or the card may be adjusted so that the bearings may be taken relative to the meridian. This card is illuminated by a small light concealed in the base. Bearings are read by means of parallel lines engraved on a

piece of heavy plate glass (Fig. 321) or parallel wires held by two arms (Fig. 322) which revolve about the compass card as the loop is turned. On ships fitted with a gyro-compass the dumb-compass card may be replaced by a live gyro-repeater. In installations of this kind the bearings are true bearings and are read from a small scale engraved on the outer edge of the gyro-repeater.

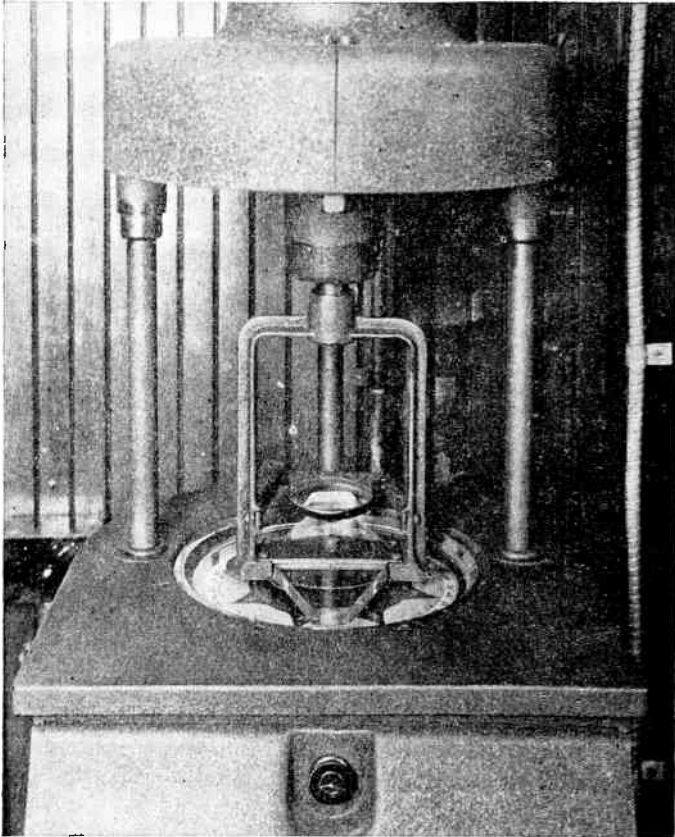


FIG. 321.—Close-up view of bearing indicator, R.M.C.A. type. (Courtesy of R.M.C.A.)

On smaller types of direction finders a 360-degree scale is provided over which revolves a pointer attached permanently to the loop shaft.

The types of marine-radio direction finders, shown in Fig. 320, usually have two pointers (colored red and white) fixed so that they point in opposite directions and at right angles to the parallel lines. These pointers are useful when it is desired to utilize the unidirectional qualities of the direction finder, as will be seen when the practical operation of the direction finder is explained.

c. The receiver utilized with a radio direction finder must be designed to give maximum sensitivity and selectivity together with simplicity of operation. When push-pull radio-frequency amplification is used, it is provided to give electrical symmetry to the loop circuit and reduce antenna effect. On radio direction finders designed for use where great accuracy is not required, push-pull radio-frequency amplification is not provided. Also, on direction finders designed for installation in seaplanes, where antenna effect is not pronounced owing to the absence of surrounding metallic objects which are grounded, push-pull radio-frequency amplification is not necessary. The expense of construction on these receivers is, therefore, proportionately less.

d. The frequency range of marine-radio direction-finding equipment generally falls within the band extending from 250 to 540 kc./s. This allows operation on the three principal direction-finder frequencies, namely, 300, 375, and 500 kc./s. Radio beacons operate on 300 kc./s., radio-compass operation is carried on between ships, and between ship and shore, on 375 kc./s., and 500 kc./s. is the distress frequency on which direction-finder operation might be necessary in locating a vessel in distress. Radio direction-finder receivers are usually calibrated for these three frequencies.

299. Grounds and Shielding.—It is the usual practice to mount direction-finder receivers in a metal case which acts as a shield and reduces antenna effect. In addition to this, any external leads from the loop or battery box (if it is a separate unit) are shielded, and all shields are grounded.

A grounded electrostatic shield is also advantageously placed between the primary and secondary of the output transformer of the receiver. This shielding is provided to reduce antenna pick-up by the telephones and connecting cord which causes residual signal.

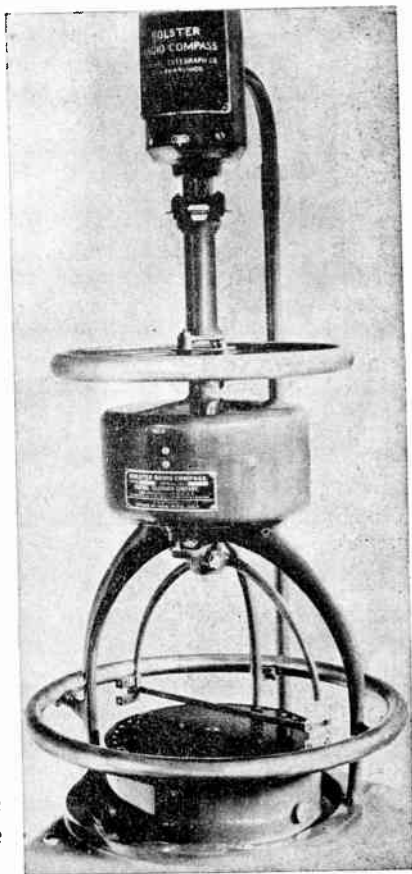


FIG. 322.—Close-up view of bearing indicator, Mackay type. (Courtesy of Mackay Radio and Telegraph Company.)

300. Signal Indication.—Signal intensity may be indicated by the visual methods or by the audio method. The latter is perhaps the most commonly used and requires ordinary radio headphones connected to the receiver to indicate maximum and minimum signal reception in the operation of the direction finder. The visual methods are less commonly employed but, nevertheless, are very practical for strong signals.

One visual system of indicating signal strength utilizes a special type of lamp containing an inert gas (usually neon gas). This lamp is connected directly in the loop circuit and its operation may be described as follows.

If the loop is rotated to a position where maximum e.m.f. is induced into it, it will be seen from previous explanations that the current will be a maximum if the loop is parallel to the line of direction of travel taken by an oncoming wave; consequently, if the loop is at right angles to the direction of travel taken by the wave, a minimum of current will flow in the loop.

If the e.m.f. induced in the loop is large enough, it will ionize the gas medium in the tube which will result in a glow. Then as the loop is rotated to a point where the minimum signal is induced into it, the lamp will not glow. Thus it will be seen that the lamp will cease to glow between two readings on the compass card, the *mean* of which will be the actual bearing.

It must be borne in mind, however, that the width of the compass sector over which the lamp is dark will depend on the strength of the transmitted signal and also the distance from the transmitting station to the ship which is taking the bearing. It has been found that a reading accurate to 2 degrees can usually be obtained over a distance of 25 miles.

An advantage of this system over the audible system is that it is possible to rotate the loop to a point so that a signal transmitted from either a land beacon or a mobile radio beacon installed on another ship will cause the lamp to glow as soon as the vessel, on which the glow lamp is installed as a part of the direction finder, comes into a danger zone, thus making it possible by this means, to have the lamp placed in a conspicuous position whereby the navigator may observe approaching danger by occasionally glancing at the lamp.

Another type of visual indicator is the meter type. The scheme here is to connect a sensitive rectifier-type voltmeter across the output of the receiver in place of, or in addition to, the ordinary headphones. The method of connecting this meter is shown in Fig. 324, which also shows the three-point switch (S_3) used to switch the meter across the headphones at will. The switch has three positions, OFF, "1" which connects a resistor in series with the indicator meter to reduce deflection if signals are too strong, and "2" for full deflection. This type of indicator should prove helpful when there are noises in the same room with the direction finder which makes listening difficult. This visual indicator will of course respond to interfering signals in the direction finder, but, under favorable conditions, coordination of the visual and aural methods of observation should improve accuracy.

301. Schematic circuit diagrams of various types of radio direction finders are shown in Figs. 323 to 325, inclusive.

a. *Ship Type*.—A type of direction finder installed on many vessels of the merchant marine and larger vessels of the Coast Guard, and around which most of this chapter has been written, is shown in Fig. 323. When this direction finder was first designed it was made to operate with

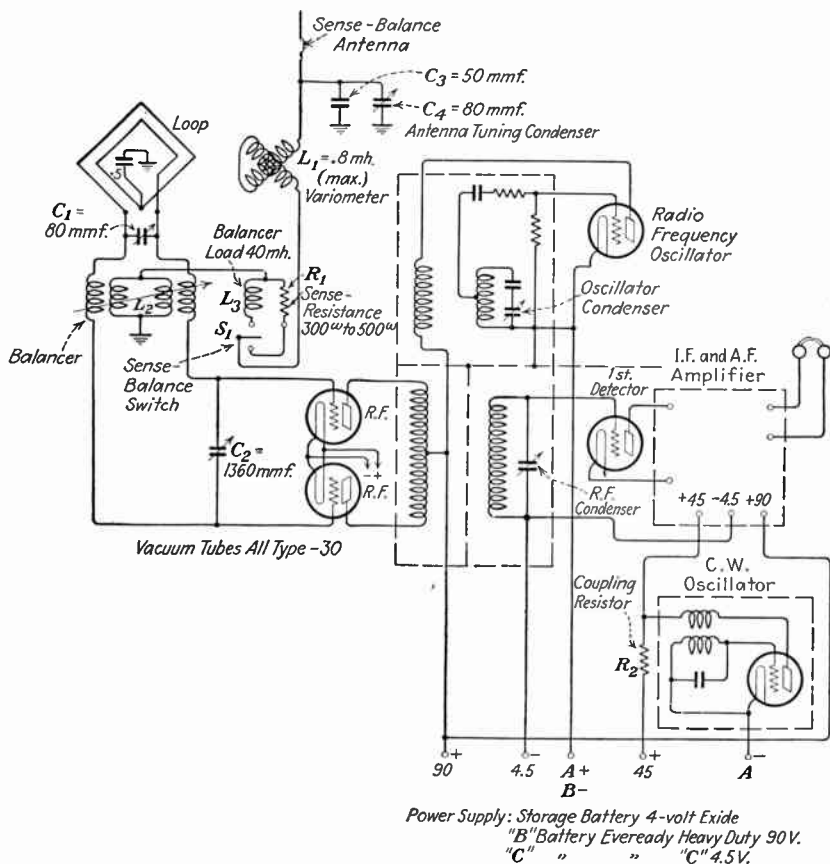


FIG. 323.—Circuit diagram of radio direction finder. (CGR-17 or -18; ER-1445.) (Courtesy of U. S. Coast Guard.)

type UX-201-A tubes utilizing a 6-volt filament storage battery, but in later models there has been substituted type -30 tubes which require only 2 volts for lighting the filament. This voltage is supplied by a 4-volt lead storage battery through a filament rheostat. The various type numbers under which this direction finder is known are accounted for largely by the type of tubes used.

b. *Coast Guard direction finder, model CGR-16*, is shown in Fig. 324. This instrument is provided for use at Coast Guard traffic stations and at important outlying base radio stations for the purpose of obtaining

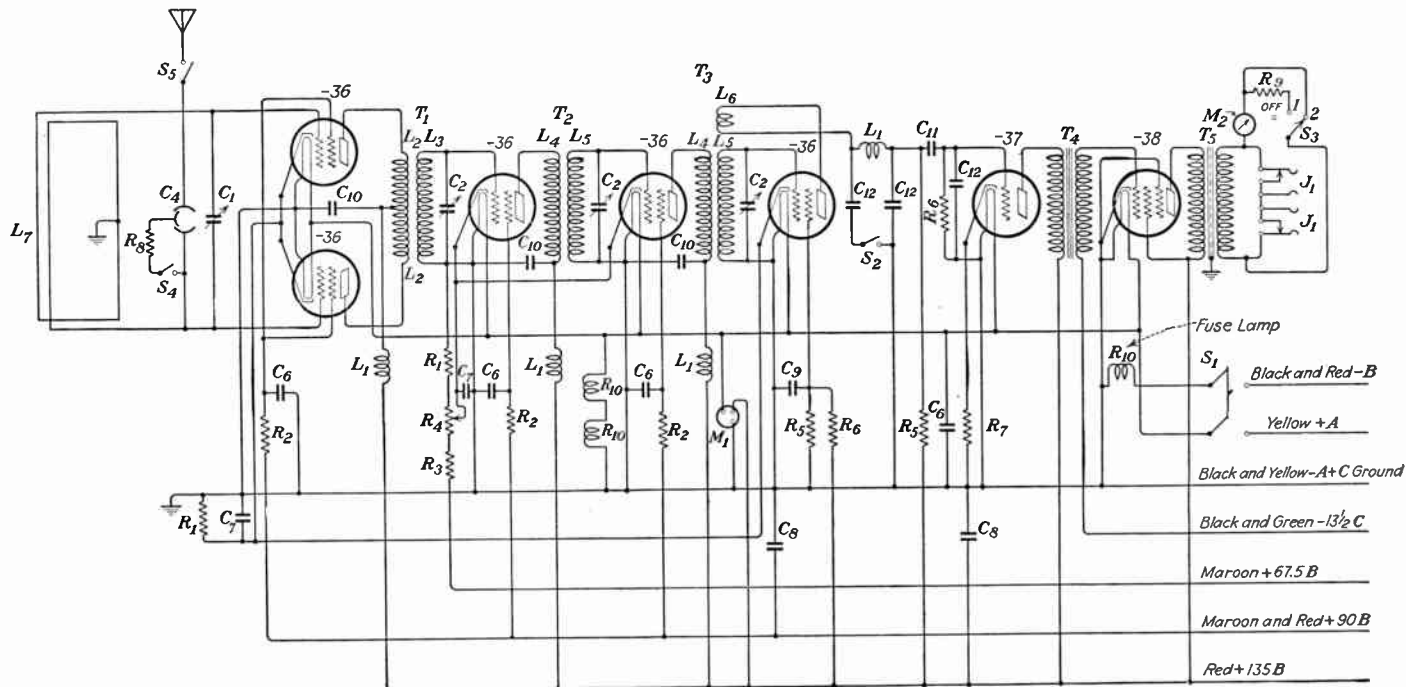


FIG. 324.—Radio direction finder, C.G. model CGR-16. (Courtesy of U. S. Coast Guard.)

PARTS OF RADIO DIRECTION FINDER, U. S. COAST GUARD, MODEL CGR-16 (See Fig. 324)

- | | | |
|--|--|--|
| C_1 Loop tuning condenser, 0.00084 mf. | L_2 One-half push-pull r.-f. transformer, primary winding part of T_1 , 154 μ h. each total 395 μ h. | R_5 Detector plate resistor, $\frac{1}{2}$ -watt size, 250,000 ohms. |
| C_2 R.F. tuning condenser 0.00042 mf. | L_3 R.F. transformer, secondary winding, part of T_1 , 1,125 μ h. | R_6 Grid resistor, first audio tube, 500,000 ohms. |
| C_3 Trimming condenser, integral with C_1 and C_2 , 0.000035 mf. (not shown). | L_4 R.F. transformer, primary winding, part of T_2 and T_3 , 150 μ h. | R_7 Cathode resistor, first audio tube, 1,500 ohms. |
| C_4 Balancer condenser, 0.00005 mf. | L_5 R.F. transformer, secondary winding, part of T_2 and T_3 , 1,125 μ h. | R_8 Sense resistor, 1-watt size with tips, 10,000 ohms. |
| C_5 Screen-grid by-pass condenser, part of C_B , 0.1 mf. | L_6 Tickler coil, part of T_3 , 50 μ h. | R_9 Visual-indicator meter resistor, $\frac{1}{2}$ -watt size, 6,000 ohms. |
| C_7 Volume-control by-pass condenser, and cathode resistor by-pass, part of C_B , 0.25 mf. | L_7 Loop winding, 435 μ h. | S_1 ON-OFF set switch. |
| C_8 135-volt and 90-volt B battery by-pass, part of C_B , 1.0 mf. | M_1 Battery voltmeter, double range, 200-8 volts d.c. | S_2 CW control switch. |
| C_B By-pass condenser block, 200 volt d.c., 1.9 mf. | M_2 Visual indicator meter, rectifier-type, a.-c. voltmeter, 3 volts, 2,000 ohms per volt, with special scale. | S_3 Visual-indicator switch, three-point tap switch. |
| C_9 Detector screen-grid by-pass condenser, 200 volts d.c., 0.25 mf. | R_1 Cathode limiting resistor, $\frac{1}{2}$ -watt size, 280 ohms. | S_4 Sense switch, high-voltage push type. |
| C_{10} Plate filter condenser, 500 volts d.c., 0.015 mf. | R_2 Screen-grid resistor, $\frac{1}{2}$ -watt size, 10,000 ohms. | S_5 Antenna sw., (single-pole single throw). |
| C_{11} Detector audio coupling condenser, Waxtite cartridge type, 400 volts, 0.005 mf. | R_3 Volume-control resistor, $\frac{1}{2}$ -watt size, 75,000 ohms. | T_1 Push-pull, 1st stage r.-f. transformer. |
| C_{12} Detector plate-filter condenser, mica 1,000 volts d.c., 0.0005 mf. | R_4 Volume-control potentiometer, 20,000 ohms. | T_2 2d stage r.-f. transformer. |
| J_1 Telephone jacks. | | T_3 3d stage r.-f. transformer. |
| L_1 Plate-filter choke, 15 mh. | | T_4 Interstage audio transformer. |
| | | T_5 Output audio transformer. |

Note: tubes, r.-f. amplifiers, type -36; detector, type -36; first audio, type -37; power-amplifier, type -38 (pentode).

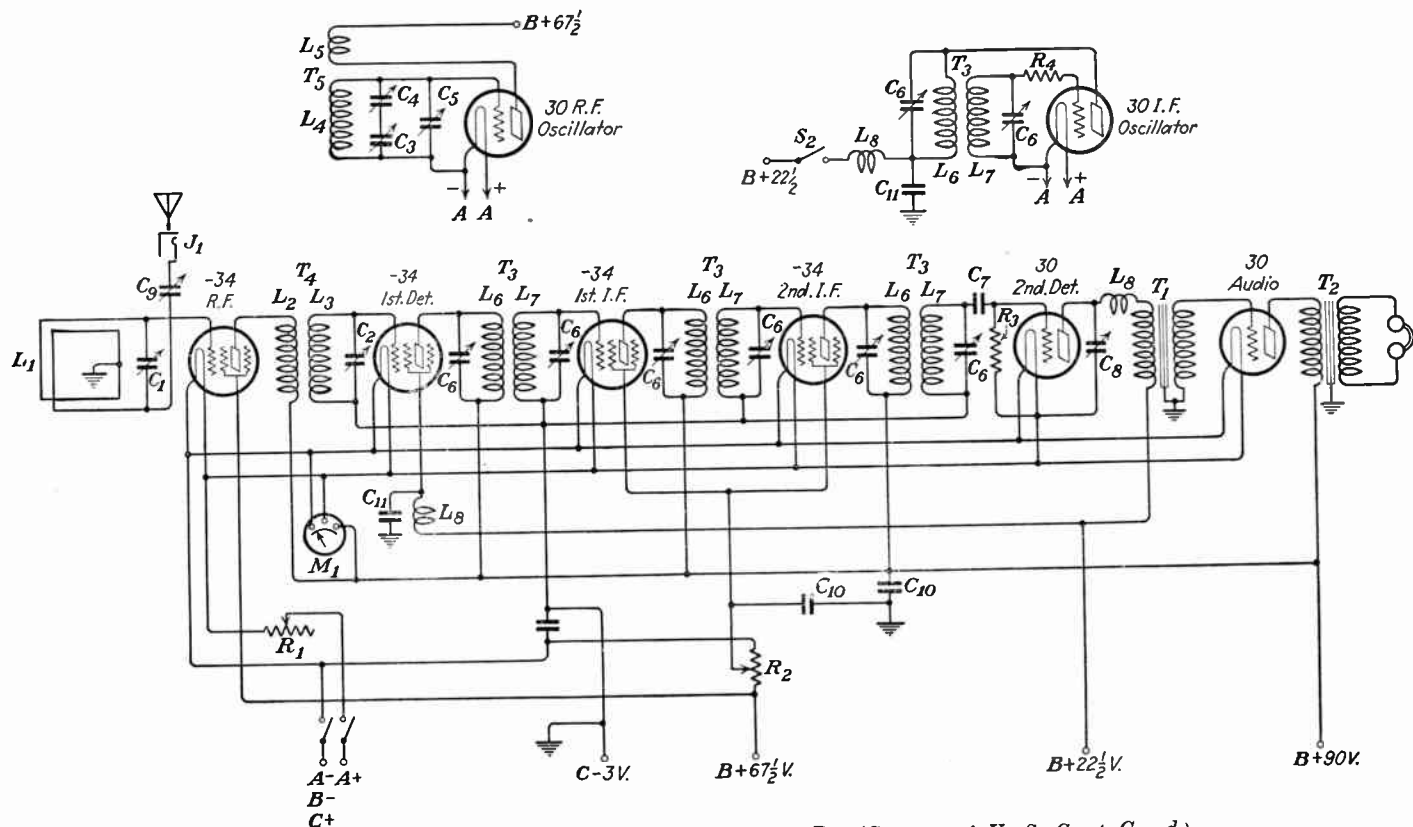


FIG. 325.—Radio direction finder; C.G. model CGR-19-D. (Courtesy of U. S. Coast Guard.)

PARTS OF RADIO DIRECTION FINDER, AIRCRAFT TYPE U. S. COAST GUARD, MODEL CGR-19-D (See Fig. 325)

- C_1 Loop tuning condenser, 0.00126 mf.
- C_2 } R.-F. tuning and r.-f. oscillator condenser, each section 0.00041 mf.
- C_3 }
- C_4 R.-F. oscillator padding condenser, 0.001 mf.
- C_5 R.-F. oscillator trimming condenser, 0.00014 mf.
- C_6 I.-F. transformer trimming condenser, 0.00014 mf.
- C_7 Second detector grid condenser, 0.00025 mf.
- C_8 Second detector plate by-pass condenser, 0.002 mf.
- C_9 Antenna adjusting condenser, 0.00070 mf.
- C_{10} By-pass condenser 250-volt non-inductive type, 2 mf.
- C_{11} Filter condenser, 0.1 mf.
- J_1 Antenna jack, open-circuit type.
- L_1 Loop winding, 350 μ h.
- L_2 R.-F. transformer, primary, 410 μ h.
- L_3 R.-F. transformer, secondary, 1,065 μ h.
- L_4 R.-F. oscillator, grid winding, 440 μ h.
- L_5 R.-F. oscillator, plate winding, 48 μ h.
- L_6 I.-F. transformer, primary and secondary windings, each 6.5 mh.
- L_7 I.-F. oscillator, primary and secondary windings, each, 5.25 mh.
- L_8 R.-F. choke, honeycomb wound, 15 mh.
- R_1 Filament rheostat, 6 ohms.
- R_2 Volume control, 500,000 ohms.
- R_3 Second detector grid, $\frac{1}{2}$ -watt size, 2 megohms.
- R_4 I.-F. oscillator-grid suppressor, $\frac{1}{4}$ -watt size, carbon, 6,000 ohms.
- S_2 CW jack switch.
- T_1 Interstage audio transformer: primary turns 5,100; secondary turns 18,000; primary resistance 2,000 ohms; secondary resistance, 9,000 ohms.
- T_2 Output audio transformer: primary turns 5,075; secondary turns 5,500; primary resistance 1,000 ohms; secondary resistance 1,500 ohms.
- T_3 I.-F. unit and intermediate-frequency oscillator unit, 175 kcs.
- T_4 R.-F. amplifying transformer.
- T_5 R.-F. oscillator transformer.
- M_1 Meter, double range 150-7.5 volts, d.c., with push button.

first-hand information concerning vessels in distress and furnishes a means for reading signals through radio interference. This equipment is not used for furnishing navigational bearings to vessels, except that certain stations specially designated by Coast Guard Headquarters may furnish bearings to Coast Guard aircraft. This direction finder is not intended for portable use and under no condition shall it be removed from its designated position unless an emergency exists warranting such action. This instrument is fitted with a meter type of visual indicator (M_2 in the diagram) which is a rectifier-type alternating-current voltmeter, 0 to 3 volts, 2,000 ohms per volt, fitted with a special scale.

c. An aircraft radio direction finder manufactured for the U. S. Coast Guard is shown schematically in Fig. 325. This type, known as model

CGR-19-D, is designed for use on heavier-than-air Coast Guard craft where a radio operator is available for taking radio bearings. The various units comprising the equipment are all mounted within easy reach of the operator and so located that the receiver may be used in conjunction with the transmitter for communication purposes if desired. The antenna connects to the circuit through a jack mounted on the receiver panel and is easily connected or disconnected as desired. The center of the loop is grounded to provide a bias for the grid of the radio-frequency amplifier tube.

A balanced radio-frequency input to the receiver is not necessary because of the absence of residual signal and deviation on a loop installed aboard an aircraft. In other respects this direction finder is similar in principle to the conventional direction finder designed for marine use, except that it is much smaller in size.

302. Calibration. *a. Requirements.*—The radio direction finder must be calibrated before it is ready for operation. This means that the direction-finder readings must be compared with true magnetic bearings and permanently recorded for the use of the operator or navigator in setting and checking the mechanical compensator, or, if such a compensator is not fitted to the direction finder, for use in converting direction-finder bearings to true bearings.

The procedure followed in calibrating a radio direction finder aboard a vessel differs in detail from the procedure followed in calibrating a direction-finder installation located ashore or on a sea-going heavier-than-air craft. Essentially, however, the method is the same in that it involves the taking of simultaneous sight and radio bearings and recording the results.

When a navigator on a vessel utilizes a radio bearing supplied by a radio direction-finder station located ashore, it is necessary that he know the exact geographical location of the shore station at which the bearing was measured. This information is to be found together with other pertinent data in the official list of Direction-Finding Stations published by the government and referred to before. For this reason, therefore, it is necessary when calibrating a radio direction-finder station located on land that the exact geographical location of the station be determined in degrees, minutes, and seconds of latitude and longitude. This requires an accurate survey map of the vicinity indicating the location of prominent objects such as tall buildings, church spires, standpipes, lighthouses, and the like, by means of which the station location and true north can be determined.¹

¹ Various methods of locating true north are described in detail in "Wireless Direction Finding and Directional Reception" by R. Keen. See bibliography; Appendix A.

The actual calibration of a radio direction finder is usually done by someone especially trained for this work. Once the direction finder has been calibrated, no one should tamper with the mechanism.

b. Calibrating a Direction Finder Which Has Been Installed on a Vessel.—The procedure required is to swing the vessel in azimuth while taking simultaneous radio and pelorus observations. These observations are recorded in two comparative columns, their differences noted, and a calibration curve of plus or minus errors against the direction-finder reading is plotted. Readings are taken from the calibration curve at every 15 degrees, the mechanical compensator (if provided) adjusted accordingly, or the variations are transferred to a corrector disk. As the accuracy of the direction finder after calibration depends directly on the accuracy of the visual observations made at the time of calibration, these observations must be made with the utmost care.

It is obvious that the radio direction finder cannot be calibrated while the vessel is in dock. This operation is usually carried out in open water (for example, off Ambrose Lightship in the New York area), and, at the time of calibration, all davits, booms, searchlights, and any other large metallic objects in the vicinity of the direction finder must be stowed in their normal position for sea and left that way throughout calibration. Changes of any extensive nature made in the location or trim of any of these surrounding objects might induce an error in the direction finder. This especially applies to large searchlights which are sometimes mounted close to the loop and which, when not in use, should be kept trimmed in their regular position. Stays and guys should be left in the condition in which they were at the time of calibration, that is, ungrounded, or grounded to the ship's hull by jumpers. To ground an ungrounded stay, or vice versa, after the direction finder has been calibrated is likely to introduce an error.

The calibration of the direction finder may be checked from time to time for accuracy by taking simultaneous pelorus and radio bearings when in the vicinity of light vessels equipped with radio beacons, or when in visual range of any vessel capable of transmitting the required signals. Signals used for calibrating purposes, or for checking calibrations, should be about 3 or 4 miles distant. At least 4 miles is preferred so that any possible error due to parallax (wave distortion) may be minimized.

Marine-type radio direction finders, such as are described in this chapter, are simple to operate, but considerable practice is required before the observer can become proficient and skillful in taking quick and accurate bearings under all conditions.

c. Calibrating a Direction Finder Installed at a Shore Station.—This operation requires that the procedure described before, for calibrating a direction finder installed on a vessel, be reversed. It is obvious that the

shore station cannot be moved and that necessarily the source of signal must be moved instead. Calibration under these conditions is usually carried out by utilizing a radio transmitter located on a small vessel which can be moved around in azimuth to the direction finder which is fixed in position. Sight and radio bearings are taken in the usual manner from true north as a reference. A transit or theodolite is required for taking the visual bearings.

For calibration purposes the transmitting signal should be modulated.

d. Calibration on Aircraft.—In general, the procedure for calibrating a radio direction finder of the manual type installed on a sea-going aircraft is similar to the method of calibrating a direction finder installed on a vessel except that the calibration must be done with the plane on a landing field and not on the water. The work must be done with the craft located well away from conducting wires or structures and with no intervening objects between transmitter and receiver. Near-by or intervening objects may produce wave-front distortion resulting in false radio bearings. The transmitter must be located so that its signal does not pass closely parallel or oblique to the shore line or across intervening objects. The effect of all transmitting and receiving antennas on deviation must be determined at the time of calibration. Those antennas found to affect the deviation must be *open* (not *grounded*) during calibration and when taking bearings thereafter.

For the transmitting station or target a medium-powered transmitter emitting modulated waves and located on a small patrol boat should be employed. The calibration should be made at, or close to, the frequency to be most used. On a properly designed direction finder the calibration will probably hold quite well over the entire frequency band, but, if closer accuracy is desired, a calibration should be made at each end of the band. The aircraft must be insulated from the ground with the tail elevated to flying level by a cart or dolly. The motive power for accomplishing the swinging may be a tractor or its equivalent.

For taking the simultaneous visual bearings, a pelorus must be securely mounted on some part of the plane where azimuth sights may be made throughout 360 degrees. The zero of the pelorus should be set on the dead-ahead position, or lubber's line.

When the head of the ship is swung into approximately the zero position, the radio observer obtains a null on the transmitter and calls out "mark." The visual observer at the pelorus notes his bearing, and both observers give their readings to the recorder. The difference between readings determines the error. Swing the tail of the ship about 5 degrees and repeat the readings. Continue with this until the ship has been turned around 360 degrees. Calibration made with the aircraft on the water must be regarded as approximate and not sufficiently accurate for navigational purposes.

Plot on a calibration chart, as shown in Fig. 306, the points observed, using the radio bearings as abscissas and the difference between pelorus and direction finder as ordinates. The shape of the correction curve may be different for the various types of aircraft owing to the difference in the locations of closed loops formed by struts and bracing in the frame.

303. Operation.—The first rule in the operation of the radio direction finder is: NEVER USE THE DIRECTION FINDER WHILE THE MAIN ANTENNA IS IN USE OR GROUNDED. To guard against accidental or unintentional violation of this rule a radio direction finder designed for use on a large vessel, where the direction finder is located in the chart room or pilot house, is provided with an interlock switch arrangement between the direction finder and an antenna-opening switch in the radio room. A signal light in the radio room notifies the radio operator when the direction finder is in use. This interlock renders the direction finder inoperative unless the antenna switch in the radio room is open.

If the radio direction finder is located in the chart room or pilot house, the navigator, when wishing to use the direction finder, directs the radio operator to open (not ground) the main antenna. Other antennas may be left connected in their normal operating condition provided they are not tuned within 1,500 kilocycles of the observed signal frequency. *Any equipment used with these antennas should not be tuned or adjusted while the direction finder is in use.* To take a bearing with the main antenna in use or grounded might result in damage to the direction finder or in an erroneous observation being made. This is due to the direct effect of the main antenna field in which the direction-finder loop is usually located. The grounding or using of the main antenna while taking a bearing is usually indicated by a strong residual signal existing in the null area and an abnormal balancer adjustment.

The actual operation of a radio direction finder is substantially in accordance with the following operations:

a. Getting the Line of Direction.—1. Start the receiver operating by inserting the telephones in the jack and operating the proper switches in accordance with the requirements for the particular set in use.

2. Tune in the source signal (signal on which bearing is to be taken) by varying tuning control on receiver to desired frequency and by orienting loop until maximum signal is heard. If volume is too great, decrease it, by decreasing filament voltage or by operating volume control, if supplied. Do not decrease volume by detuning receiver as this may result in erroneous bearings.

3. Swing the loop approximately 90 degrees from the position of maximum signal which should bring it to the area of minimum signal audibility. If a parallel-line indicator is provided, observe from proper end. This is the end which was used during calibration. On the plate-glass type it is indicated by an arrowhead. Swing the loop back and forth over this restricted area, while at the same time adjusting the control marked BALANCE, until a point is found where the signal strength is zero or nearly zero. Set the loop at the center of this null area. Read the desired bearing

from the indicator. If the observed signal is from a distant source and is weak, the null will be broad. In this situation increase the filament voltage to normal, or bring up the volume by operation of the volume control if heater-type tubes are used. Conversely, a very strong signal from a near-by transmitter will give an exceedingly sharp null which may be confined to a fraction of a degree in width. On exceptionally loud signals, it may be necessary to decrease the volume considerably in order to discern the area of minimum signal intensity. If filament-type tubes are used, care must be taken not to decrease the filament voltage so low that the receiver becomes inoperative. If the null extends over a considerable range, it will be necessary to note two points, one on each side of the silent area, where the signal becomes of equal audibility or intensity, taking the *mean* of these two points as the bearing.

On some direction finders, due to local effects and the characteristic of the signal under observation, it may be impossible to obtain an absolute null as the indicator is swung through the minimum area. When this condition exists the bearing will be taken from the estimated center of the area of minimum signal intensity. This condition is shown in Fig. 312.

The quality and definition of the null will vary on a given signal, as it is observed from various headings. This is due to the effect on the incoming signal by local metallic agents such as stays, guys and riggings. This effect has been explained in detail under Sec. 292 *b* on Reducing Deviation. It may now be said, however, that at times the residual quality of an observed signal may be too strong for complete extinction.

The radio bearing indicated by the point of minimum signal on the scale is either the true bearing from the observing station or its reciprocal. For example, if a minimum signal is observed when the indicator points to 60 degrees, another minimum signal will be observed when the indicator points to 240 degrees. Unless the observer knows positively on which side of the compass the transmitted signal lies, he would have no way of telling which of these bearings was the true bearing. This difficulty is eliminated by utilizing the unilateral characteristics of the loop when used in conjunction with a non-directional antenna and is called *getting a sense bearing*. This operation follows next in order.

b. Getting the Sense of Direction.—1. Swing loop approximately 90 degrees from point of null position, and throw switch to *sense* position (or push *SENSE* button, if supplied instead of switch), noting signal strength. Then turn loop approximately 180 degrees (sense switch or button still in) and note signal strength. It will be noted that the signal is stronger in one 90-degree (from null point) position than in the other. On the conventional type of marine direction finder found on large vessels the quadrant in which the true bearing lies will be indicated by the position of a pointer (usually red) when the loudest or maximum signal strength is obtained with the sense switch on.

Another method of obtaining the sense of direction is as follows: The true bearing is indicated by the way in which the loop is turned for maximum signal. If the position of strongest signal occurs when the front side of the indicator moves in a clockwise direction, the bearing as indicated on the indicator scale is correct. Otherwise, the bearing is the reciprocal, the loop should be turned 180 degrees, and the correct bearing determined by a new point of minimum signal which should be the reciprocal of the first bearing observed.

In obtaining sense bearings the sense effect will be more apparent if the sense connection within the circuit is made intermittently, rather than closed permanently, while swinging the loop through the area of maximum signal intensities. This can be done by rotating the knob controlling the sense switch to the sense position several times during the sense operation or by pushing and releasing several times a sense button provided on some types of direction finders for this purpose.

2. The true bearing, as indicated by the proper null point, should have added or subtracted to or from it the necessary correction factor unless this operation is automatically done by a mechanical compensator.

3. If bearings are to be taken on *CW* signals, the *CW* control switch must be closed. On direction-finder equipments provided with superheterodyne receivers this *CW* switch starts the *CW* oscillator which beats against the incoming signal making it audible. On direction-finder equipments not fitted with superheterodyne receivers this switch closes a feed-back circuit on the detector causing it to operate as a regenerative detector and thus making it capable of receiving *CW* signals (see Wiring Diagram, Coast Guard model CGR-16, Fig. 324, S_2).

4. If the direction finder is located in the chart room or pilot house and used by the navigator directly, he should notify the radio room when he is finished with the direction finder so that the main antenna may be placed back in service.

304. Fixing the Position.—The operation of fixing the position of a vessel or aircraft, after the radio bearings have been taken, is called plotting a "fix" and is strictly a navigational function. It is not generally done by the radio operator but is performed by the captain or another navigating officer.

Figure 326 illustrates a method by which a ship approaching New York harbor might fix its position. In this method two bearings are taken, one from Ambrose Channel Lightship, and the other from Fire Island Lightship. In the illustration it is found that Ambrose Channel Lightship is 20 degrees off the port bow (to the left) and that Fire Island Lightship lies 40 degrees off the starboard bow (to the right). By applying navigational deductions and plotting the lines on a chart, the position of the observing station is at the point where these lines cross. A third bearing might be taken, as for example from the Sea Girt Light Station, as shown in the diagram, to increase the accuracy of the "fix."

Another method is to take two bearings from a single source, this taken with the distance traveled may be solved as a triangle problem by trigonometry. For example, in the triangle in Fig. 327, let A be the source of signal or the transmitting beacon; let B be the position of the ship when the first bearing was taken, let C be the position of the ship when the second bearing is taken; a is the distance between B and C or the miles logged, and b is the distance from the source A to the observer C . Then, $b = \frac{a \cdot \sin B}{\sin A}$. The entire problem is worked with the aid of a navigation chart and the exact position of the observer easily marked off.

In a similar manner radio direction-finder stations located on the shore can take simultaneous bearings on a vessel equipped with a transmitter which bearings can be plotted on a chart on which the position of the vessel will be indicated by the point of intersection of the lines if the triangulation method is used, or by trigonometrical deduction if the second method is used.

If three or more bearings are taken on a ship by three or more well-separated shore stations, the lines will in nearly every case form a small

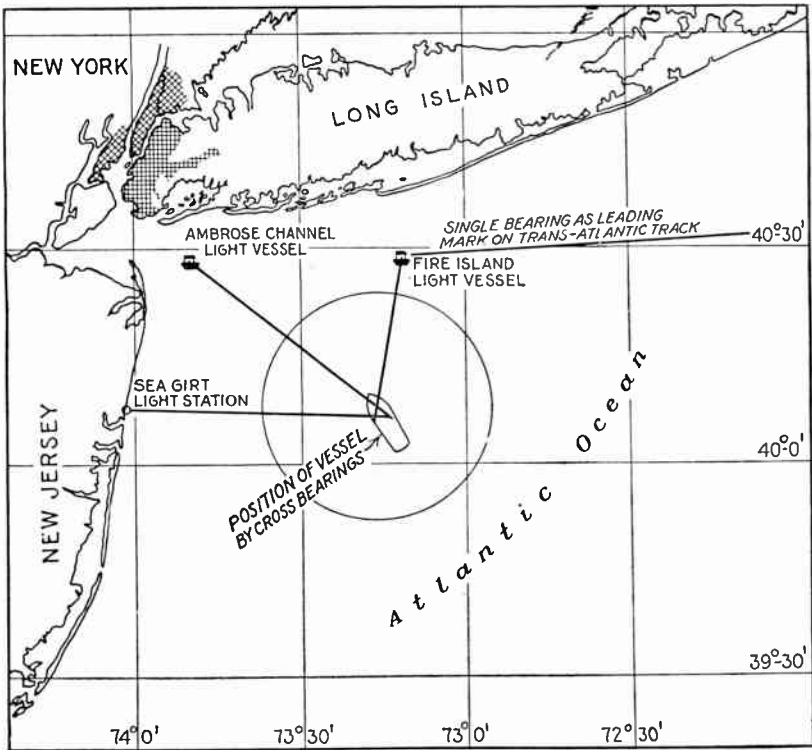


FIG. 326.—Method of plotting position from radio bearings.

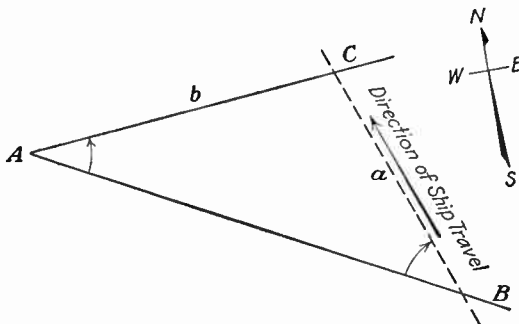


FIG. 327.—Position plotting by two bearings from a single source.

triangle by their points of intersection. The center of this triangle is regarded as the location of the vessel. The only direction-finder stations

located on the shore are those maintained by the U. S. Navy and the U. S. Coast Guard; the latter does not furnish bearings to ships as a regular procedure unless distress or emergency conditions are involved. At all these stations only operators, especially trained for this duty, furnish navigational bearings.

305. Maintenance. *a. General.*—Routine inspection of the direction-finder apparatus should be made at least once each week. During this inspection measure the voltage of the A and B batteries with the set turned on. Unless the B battery is on trickle charge, measure the gravity of each cell, and charge if necessary. B batteries should be replaced when the voltage has dropped 15 or 20 volts for the entire bank. C batteries should be replaced at the same time.

Test receiver for response to signals. Low sensitivity may be due to one or more defective tubes which may be detected by replacing each tube independently until signal response is normal.

Inspect the lubrication every 3 months and if required add a small amount of grease at the proper points.

Rotate loop by a series of jerks while the receiver is in operation to test slip rings for noisy contact. If these contacts are found to be noisy, clean with canvas or fine sandpaper.

When a vessel is in port and there are probabilities of the ship's power being shut off, it is advisable to disconnect the filament battery in the event of a polarity reversal when the power is again put on. It is recommended here that all operators should test for proper polarity before reconnecting, in order to avoid the possibility of ruining the storage battery. Serious consequences in a case of this kind may result.

b. Aircraft Direction-finder Maintenance.—In addition to the points of general maintenance outlined above, the direction finder installed on an aircraft deserves special consideration. Check the receiver for operation by listening to any available signals. The *CW* switch gives a squeal on broadcast or other *CW* carriers. Turn the loop to determine if nulls are obtainable on the various signals heard. Do not assume that directions of signal sources heard while the ship is in the hangar are correct. Usually, under these conditions, there will be large errors, and, possibly, all signals will come in on approximately the same bearing. This condition is due to the metal framing in or about the hangar. Observed signal intensities will be found much lower when the ship is in the hangar than when it is in the open.

If changes are made at any time in the framing of the wings, or if wiring is added to or removed from the ship, the calibration of the direction finder must be checked. When doing this, it is necessary to check only on a few points, especially those where the greatest errors occur.

306. Troubles and Remedies.—*a.* If the filament voltmeter does not read, check filament battery.

b. If the filament voltmeter reads correctly, but the receiver sounds dead, inspect as follows:

1. Check all B and C battery connections.
2. Look for poor tube.
3. Check wiring of receiver with aid of diagram.
4. Inspect contacts at collector rings.
5. See that no plate-circuit fuses or fuse lamps are blown.

c. If receiver is noisy, look for loose connection at battery connections and at collector rings. Noise is very often caused by defective or weak B batteries or by a poor phone-jack contact.

d. The entire absence of directional effect may be caused by the loop circuit becoming grounded. Moisture deposit, even if very slight, in the loop-collector-ring housing may ground the loop circuit. An open-loop circuit will also result in no directional effect.

e. Troubles that might develop in the wiring of the receiver can be located by tracing out the leads as shown in the wiring diagram of the set, supplied with the equipment. Any special directions given in the instructions applying to the particular equipment should be explicitly followed.

III. RADIO AIDS TO OVERLAND AIR NAVIGATION

307. Radio-range Beacons.—Aeronautical radio-range systems¹ have used loops as transmitting antennas with fair success in the directive transmission of radio waves over short distances. The original installations over American airways utilized these loop transmitting antennas. The wave beams were particularly steady in their directive properties during the day using loop antennas but unfortunately were relatively inaccurate at night owing to phase displacements and shifts due to wave reflections of an uneven character from the Kennelly-Heaviside layer. This variation in wave strength is called the *night effect* and has already been fully explained.

The night effect is particularly active in mountainous areas and even over flat ground when the distances exceed 30 miles or more from the beacon. A most important development for the elimination of night effect has been the T-L antenna system which was developed by the Bureau of Air Commerce, U. S. Department of Commerce, and which

¹ Full information concerning radio aids to air navigation including maps showing aeronautical radio-range beacons, airways communication stations, and intermediate airway radio stations in the United States may be found in *Airways Bulletin* 1 (or its superseding publication) which may be secured by interested persons from the Bureau of Air Commerce, U. S. Department of Commerce, Washington, D. C.

will be explained in detail later. This system provides a satisfactory beacon course throughout the entire distance range of the beacon with negligible night effect. As loop transmitting antennas may sometimes be used for such purposes as runway-localizing beacons for landing purposes where the range is only about 10 miles and night effect is not involved, its theoretical operation will be explained. The explanation which follows has to do with the loop as used in these radio-range beacons and the results of night effect.

308. Action of Loop Transmitting Antennas.—The fundamental principle of directive transmission is simply one in which the transmitted energy is trajected into four beams or paths representing the four major points of a compass, namely: north, south, east, and west. A plane flying in one of these paths or beams is so equipped as to receive the energy by either a visual or an aural electromagnetic system of wave indication. This incorporates the general principles of receiving circuits together with special aural and visual indicators. The special adaptation of radio receivers to radio-beacon reception was covered in Sec. 291 and will be further treated in Sec. 311.

If four waves are directionally transmitted into the north, south, east, and west directions by a system of parabolic reflection antennas, loop antennas, or vertical radiators, and each wave differentiated from the other by a standardized code-signal identification, the flight course may be readily determined. This is known as the *aural method* of beam identification and utilizes the Morse code of dots and dashes. On the other hand, if the four waves are transmitted at two different modulated or *tone* frequencies somewhere in the low-frequency range between 60 and 100 cycles, these frequencies, when amplified, may be made to vibrate small metallic reeds or ribbons on a visual indicator. This is known as the *visual method* of beam identification.

The transmission of beams in the radio-range beacon system was first carried on by two loops placed at right angles to each other as illustrated in Fig. 328(a). These two loops when energized by a vacuum-tube transmitter produce two figure-of-eight field patterns in conformity with the general theory of loop antennas previously explained. These two patterns produce a directional characteristic in four directions, as illustrated, and represent the four beacon courses. This condition is ideally represented during the daytime periods, since the waves do not penetrate into the Kennelly-Heaviside layer and consequently are not subject to reflections. In the night periods, however, between sunset and sunrise, these waves are reflected back to earth so that the ground and sky waves meet at some remote point from the transmitter. This is particularly true during winter nights because of the greater rate of change in the height of the reflecting layer. In other words, as the distance from the transmitter increases, there is a greater reduction of the ground wave, and, consequently, the ratio of the horizontal (sky) to the vertical (ground) signal component at the receiving end is much greater. This difficulty may be somewhat reduced at the receiving end by the use of a vertical antenna, which is much more effective to the low-angle radiation along the earth (vertical component) than to the reflected or high-angle radiation from the Kennelly-Heaviside layer.

The effect of the reflecting layer upon the received signal is therefore the same as if the two loops at the transmitting station were being oriented in different directions, that is to say, as if the beam were being continuously shifted. The effect at the receiving end is the same as if the loops' figure-of-eight patterns had actually been changed from the position (a) to the position (b) or (c) in Fig. 328. Obviously, the

results at the receiving end would show "off-course" indications anywhere between position (a) and position (b) or (c). For example, it will be noted from the diagrams that the direction of the beacon wave or course will be at the points of the intersection of the two figure-of-eight patterns. Hence, if the direction of the four courses is represented by the letters A, B, C, and D as in the first position, the direction will effectively change at the receiving end just as if the loops were changed to the second position, and the beacon wave will be in the direction of the letters A¹, B¹, C¹, and

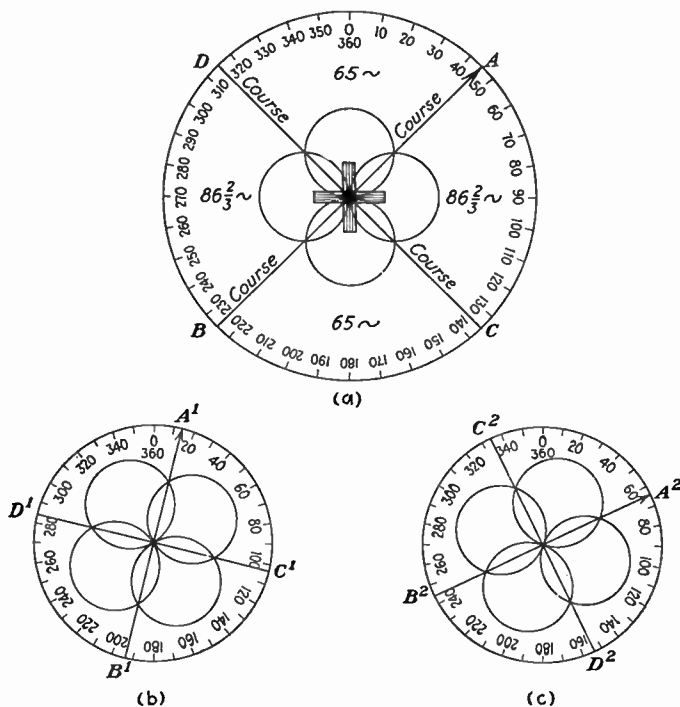


FIG. 328.—Effect of Kennelly-Heaviside layer on loop's figure-of-eight pattern.

D¹ or A², B², C², and D². It must be borne in mind that the loops themselves at the transmitting point have not oriented physically but are merely here placed in this position diagrammatically to show the results that may be expected at the receiving end because of the so-called *night effect*.

It is apparent therefore that the night effects may be reduced to only a negligible degree by the use of a radiation system in which the horizontal-radiation component is decreased to a minimum degree such as is obtained with the transmission-line type of antenna next to be explained.

A schematic diagram of a type of radio-range beacon system of loop antennas is shown in Fig. 329.

309. Transmission-line (T-L) Antenna System.—The arrangement of a transmission-line antenna system is graphically shown in Fig. 330.

The antenna comprises four vertical antennas which are placed at the corners of a square. This arrangement supplements the loop antennas used in the earlier beacon systems. The four vertical radiators are excited from a tube transmitter through a loop or goniometer system, properly phased, and a transmission line. All units are completely shielded to prevent the radiation of waves from any of the component parts. With the proper adjustments of all circuits it is found, for example, that, if the normal range of the antenna and transmitter is 100 miles, practically no night effect is detected at this distance with the transmission-line system, while, with the double-loop antenna and a transmitter, having the same transmission range, the night effect is noticeable at almost any point beyond 30 miles. Readings become extremely erratic and course indications are generally inaccurate. The variations when using the double loop beyond the 30-mile point average ± 30 degrees, while, with the transmission-line antenna, the variations are less than 4 degrees.

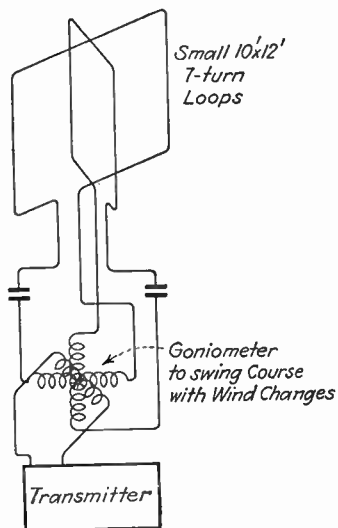


FIG. 329.—Radio-range beacon system of loop antennas.

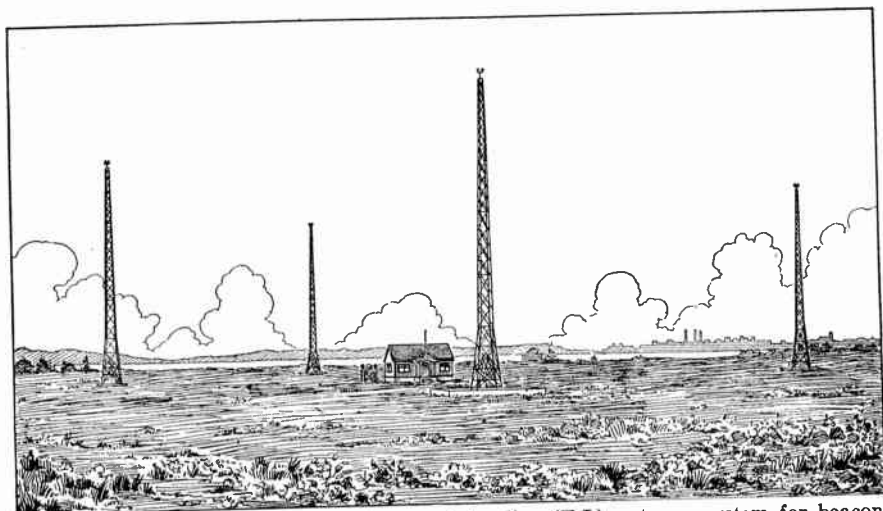


FIG. 330.—Topographical layout of transmission-line (T-L) antenna system for beacon system. (Courtesy of Air Commerce Bulletin.)

Other advantages obtained by the transmission-line antenna system are the simplification of the problem of adjusting the four beacon courses

to the arbitrary angles required by the radio waves radiating from it. Rigid adjustments are made to assure the proper adjustments of the phases and for continued stability. This antenna also provides better performance of both radio-telephone and beacon signals.

The development of the transmission-line antenna for low-angle radiation and directional transmission was based upon the fundamental experimenting of Adcock and Eekersley with vertical radiators such as are illustrated in Fig. 331. In this system the transmitting circuit output feeds energy into a transmission line in such a manner that the two vertical antennas receive voltages 180 degrees out of phase with each other, so that the current flowing in each vertical antenna is in an opposite direction to that in the other antenna. This results in a magnetic

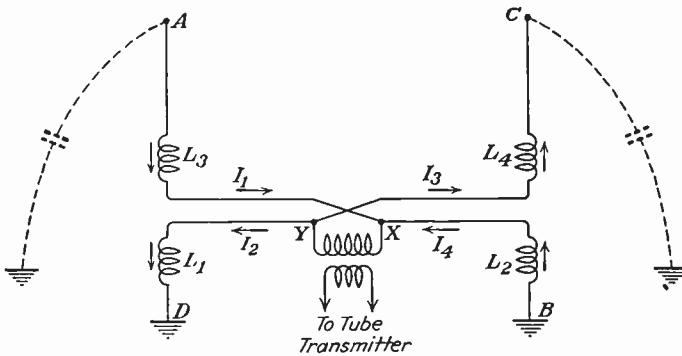


FIG. 331.—Adcock-Eekersley experiment with vertical radiators.

disturbance about each antenna equivalent to the disturbance produced by the vertical wires of a loop antenna, but, since no horizontal portion is present as in the loop, little or no energy is radiated in the horizontal plane. The only portion of the two vertical-radiator systems in which horizontal radiation might take place is that of the two parallel feeding wires or transmission lines, but, since the current is flowing in opposite directions in these two wires, the magnetic fields about them will cancel each other. However, to insure the complete cancellation of the magnetic field to prevent horizontal radiation, the currents in the two transmission lines must be of equal values, that is, I_1 must be equal to I_2 , and I_3 must be equal to I_4 . To satisfy this condition, critical adjustments of the impedances must be made so that the portion of the circuit from X to A must have an impedance equal to that of the portion from X to B. Similarly, the impedance from the point Y to C must be equal to that between the points Y and D. This condition may be obtained by inserting balancing impedances L_1 and L_2 in the vertical portions of the ground leads B and D. However, the difficulty encountered in obtaining a critical impedance match and the resultant increase of the circuit resistance due to this impedance reduce the efficiency of the entire

system to a marked degree. In the vertical-radiator systems each antenna is tuned in the conventional manner by series-loading inductances L_3 and L_4 , respectively.

The final adoption of the Adcock principle to the transmission-line system was made possible by the development of efficient transmission lines in which the maximum degree of power, developed by the transmitter, could be transferred directly to the radiating system with a very small degree of power loss. Figure 332 illustrates the electrical arrangement of the transmission-line antenna system for four-point directive transmission.

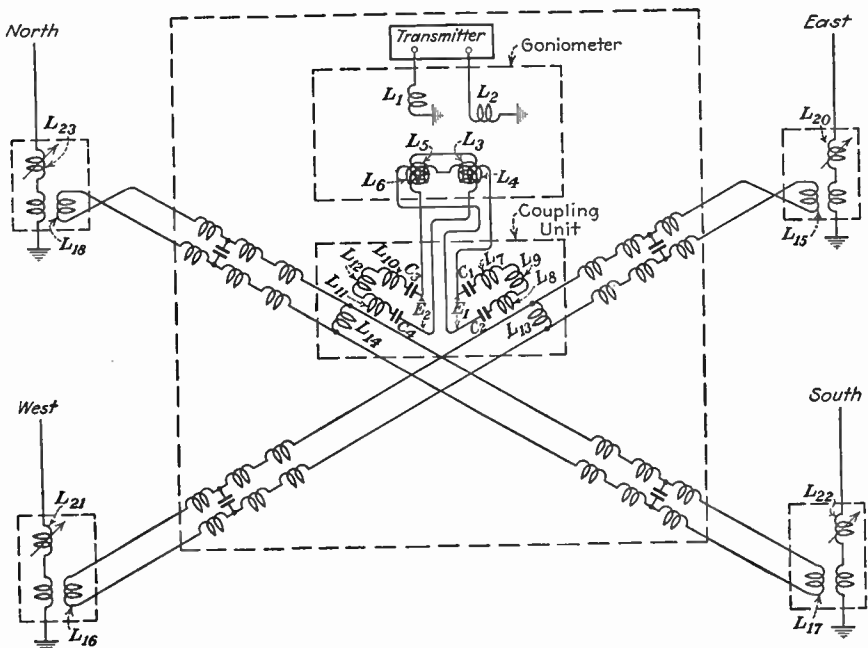


FIG. 332.—Circuit diagram of four-point directive transmission-line antenna system.

Theoretical Operation.—When the vacuum-tube transmitter is set into operation, the currents circulating through the plate-load inductances L_1 and L_2 set up varying magnetic fields about these windings. These varying fields will induce an e.m.f. across the two windings of the goniometer L_3 and L_6 which supplies the excitation for the two series-resonant circuits $L_4, L_6, C_1, C_2, L_7, L_8, L_9$, and $L_3, L_6, C_3, C_4, L_{10}, L_{11}, L_{12}$. The voltages developed across each series combination at the points E_1 and E_2 , respectively, are analogous to two simple series-resonant circuits in which the e.m.fs. across each generator are 180 degrees out of phase. Hence the voltages developed across the secondary windings of L_{13} and L_{14} in Fig. 332 will also be 180 degrees out of phase with respect to each other, and consequently the current flow in each vertical antenna will be in opposite directions. To insure that this current-phase difference prevails, the connection of the secondary L_{13} to the transmission line is reversed with respect to the connections of the other secondary L_{14} as illustrated. Now since these secondaries are to feed energy into the antenna through the four

antenna primary coupling coils L_{15} , L_{16} , L_{17} , and L_{18} , in accordance with the general theory of transmission lines, each coil must be carefully adjusted to the proper impedance of the line so that all energy transferred from the secondaries will be absorbed by the primary coupling coils. In other words, the impedance looking into the line must be matched by an equal impedance looking out of the line in order that all the power transmitted by the line be absorbed by the primary antenna coupling coils L_{15} , L_{16} , L_{17} , and L_{18} to prevent line reflections or standing waves which would result in radiations from the transmission line and seriously affect the desired four directional patterns of the vertical radiators. The impedance of the lines used in the transmission-line system is 80 ohms. In order to fulfill the condition of impedance matching, each secondary is calculated to have an impedance of 80 ohms which

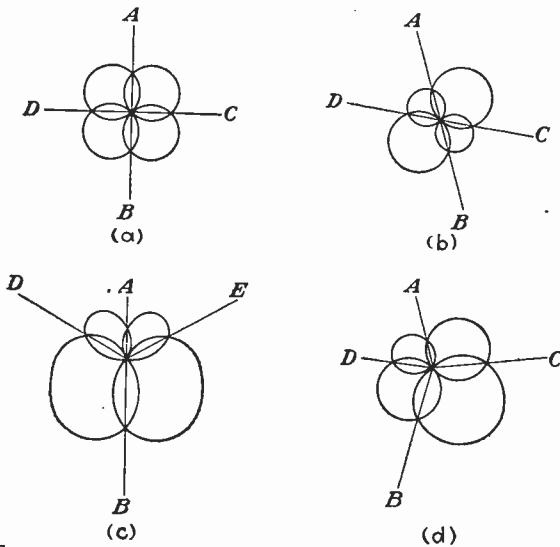


FIG. 333.—Field pattern variations, transmission-line system.

functions into the transmission line and a primary antenna-coupling impedance of the same value. When the lines are properly adjusted to these values, the radio-frequency ammeters in each line will read exactly the same. With the proper adjustment of the antenna circuit to resonance by the variation of the antenna-tuning inductances L_{20} , L_{21} , L_{22} , and L_{23} , the maximum power will be absorbed and radiated by the four vertical-antenna systems. This condition will be indicated by a maximum reading on the antenna ammeters, but great care must be taken when the antenna is tuned so that the impedance reflected into the primary coupling coils does not upset the impedance match of those windings. In other words the impedance looking into the primary windings L_{15} , L_{16} , L_{17} , and L_{18} (80 ohms each) must not be altered by the effect of the magnetic coupling reactions from the secondaries back into the primaries.

If all of these conditions are properly fulfilled, the proper time-phase angle of 180 degrees may be obtained between the currents in the two vertical antennas of each pair, and a true figure-of-eight pattern is obtained.

Figure 333 illustrates the various patterns produced by varying either the time-phase angle or the intensity of the currents flowing in the four vertical antennas. In this manner any one of the four conditions of course alignment shown may be obtained:

The normal space pattern for a given beacon in which two pairs of 180-degree courses *AB* and *CD* are produced at right angles is shown at *a*.

The condition in which the two courses are no longer at right angles or 90 degrees is shown at *b*.

The condition in which the 180-degree relationship of one pair of courses is changed is shown at *c*.

The condition in which the courses *AB* and *CD* are changed from the 180-degree relationship is shown at *d*.

The last two conditions *c* and *d* are obtained by shifting the time-phase angle between the currents in the two antennas of a given pair by inserting a variable reactance (X_L or X_C) at the output of L_{13} or L_{14} (Fig. 332) in series with either one of the transmission lines which feed the two antennas. The second condition, *b*, is obtained simply by reducing the energy in one of the transmission lines by introducing a resistance at the amplifier output of that particular branch.

It is apparent, therefore, that the time-phase angle of the beacon courses into space can be quickly varied so that a rapid phase check and alteration can be made if desired. This is accomplished by placing a sensitive thermocouple in one lead of a given pair of antennas and removing the excitation from the transmission line into this particular antenna system. Now, if the other pair of antennas are excited by their transmission lines, some current will be induced into the unexcited pair since they represent tuned or resonant absorbing circuits. Consequently, if the current induced into these antennas is zero, it is a true indication that the phase angle between the two pairs must be exactly 180 degrees. On the other hand, however, if the indicator shows a flow of current, it is apparent that the phase angle lies somewhere between +180 degrees and -180 degrees. The exact amount of 180-degree phase departure between antennas may thus be accurately determined by a meter which reads the actual angular variation in degrees.

This instrument may be a milliammeter located in the transmitter room and energized by a separate pair of underground triple-shielded transmission lines feeding from the thermocouple unit located in the antenna system under test. These leads must be thoroughly shielded, by-passed, and provided with radio-frequency choke coils at both ends to prevent radio-frequency pick-up by the meter leads.

310. Radio Beacon-signal Methods. *a. The Aural System.*—Figure 334(a) illustrates the principle of the aural type of radio-range beacon which uses the Morse code letters for identification. The signals transmitted from this system are the Morse letter N represented by the code signal - · and the letter A represented by the code signal · -. These two signals are interlocked so that when they are transmitted the first sound heard will be that of the dash in the letter N signal, then the dot of the letter A signal, followed by the dot of the letter N signal and finally the dash of the letter A signal.

When the craft is moving in the correct course, or along the "on-course" or "equisignal" zone, the two code signals merge into one long dash to form the letter T in the Morse code. In other words, if the craft is within the "on-course" zone, then the double signal will be heard in the telephones with fairly equal intensity and when the craft moves into the "off-zone" area, one signal will be louder than the other. That is to say, if the craft is in the zone area of the Morse signal N, then that signal will be the only one audible, while, if the craft is in the A-signal

area, then that will be the only signal audible. Thus, when both signals are audible, as a long dash, it is a true indication that the correct course is being followed. By maneuvering the craft in one direction or the other when "off course" the "on-course" heading will be reached in which both signals interlock to form the single or long-dash signal T. The zone in which either the N or the A signal predominates is generally referred to as the N or A quadrant, respectively. In order to establish a

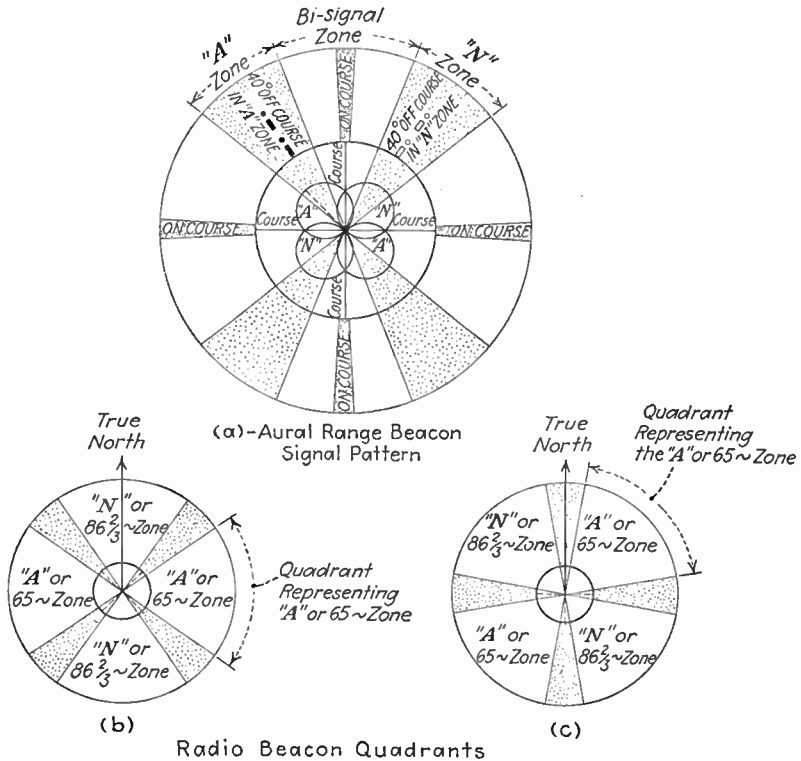


FIG. 334.—Aural system, radio beacon.

definite direction of the on- and off-course indications with respect to the true north, the N characteristic has been assigned to the quadrant through which the true north line from the transmitting station passes. If the true north line coincides with the center of a zone in which an equal intensity signal is obtained (the center of an equisignal zone) then the adjacent quadrant on the west is the N quadrant. These conditions are illustrated by Fig. 334(b) and (c).

b. The Visual System.—In this system the beacon transmitter is modulated to produce two different notes or tones of 65- and $86\frac{2}{3}$ -cycle note corresponding to the A and N signal. In other words, the radio-

range quadrants in the visual system are expressed in 65 and $86\frac{2}{3}$ cycles to substitute for the two letters used in the aural-beacon system.

The transmitted signals operate a course indicator composed of vibrating reeds, or a pointer type of instrument. In the former, two reeds are used, placed parallel to each other in a small metal case with tips painted white, both being visible through a glass panel. A nameplate is provided to indicate the frequency to which each reed is resonant. These reeds are composed of a magnetic material situated close to an electromagnet so that when a signal is passed through the receiver, the modulated component of the radio wave, when rectified by the vacuum-tube detector, will pass through the electromagnet and cause magnetic variations. Since these variations may be of either 65 or $86\frac{2}{3}$ cycles, or both, depending upon the position of the craft in the beacon course, the reeds will vibrate in direct proportion to the strength of the received signal owing to their mechanical resonance to the electromagnetic variations. In other words, if the craft is OFF COURSE, either one or the other of these reeds will vibrate depending upon which of the two frequencies is prevalent. If the craft is in the ON-COURSE zone, both reeds will vibrate equally. Hence, by noting the amplitude of the rapid vertical vibrations of the reeds the pilot may determine whether he is approaching the on- or off-course position.

With the pointer type of indicator, which is actuated by a reed converter using a pair of reeds similar to those of the reed indicator, visual indication is provided by a microammeter, the course direction being marked by the position of a pointer on a "course" or "off-course" scale. A reversing switch is provided to insure that the deflection of the pointer is in the same direction as the deviation of the craft from its course.

c. Directional Dot Transmissions.—The directional effect of a later system which utilizes the dot method of transmissions is illustrated in Fig. 335. In this system a 1-dot signal is sent to the west, 2 dots to the east, 3 dots to the north, and 4 dots to the south. This method obviates the uncertainty as to either course or direction. Experiments have shown that this system is most desirable if the 1- and 2-dot signals are transmitted successively, repeated once, and then followed by the 3- and 4-dot signals in the same order. This will enable the pilot to determine which of the 1- and 2-dot signals, and which of the 3- and 4-dot signals, are of greatest intensity. Hence, if a craft is lost in an A sector and the 1-dot signal is of great intensity, it is apparent that the craft is in the sector west of the beacon and not east of it where the 2 dots would be of loud intensity. Similarly, if the 1-dot signal is heard, but not the 2 dots, then the craft is on either course D or E. However, when the 3- and 4-dot signals are transmitted, if only the 3-dot signal is heard, the

craft is following course E. If, on the other hand, the 4-dot signal is heard, and not the 3-dot signal, the craft is on course D.

This method also permits identification of the zero-zone signal when a craft is passing over a beacon so that a pilot may determine whether or not the craft has passed over the beacon. No signal is heard directly above the transmitter if vertical radiators are used such as in the new transmission-line antenna system. For example, if a craft is flying towards the beacon located at *O* in Fig. 335 on course *OF*, the 2- and 3-dot signals will be of loud intensity, but when passing over *O* and into the course *OD* southwest of the beacon, the 1- and 4-dot signals will become loud.

With this system of transmission, sharply directive signals will be radiated in four different directions when the currents in the two opposite antennas are properly phased such as in the transmission-line antenna system. The dot method of transmission is also applicable as a quadrant identification for the aural, visual, or combined aural and visual systems. For example, when combined with the aural-beacon system, the directional identification dot signals are transmitted during the period in which the A- and N-course signals are interrupted. That is to say, the 1- and 2-dot signals may be transmitted during one of the interruptions, and the 3- and 4-dot signals transmitted during the next interruption. Approximately 2 sec. are required for single- and double-dot transmission, and approximately 3 sec. for the 3- and 4-dot transmissions. Each pair of signals should require a total of 10 sec. every 2 or 3 minutes.

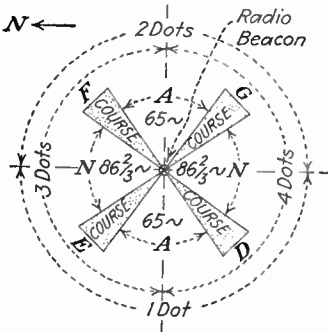


FIG. 335.—Directional dot system, radio beacon.

When combined with the visual-beacon system, the two-tone course frequencies of 65 and $86\frac{2}{3}$ cycles are interrupted only for the time interval of each dot of the group of identification signals. During the interim between the dots, the course signal is on again. The short duration of the dot signals is required in order that the reeds on the visual indicator shall not have sufficient time to drop more than one-third of their full-amplitude position.

311. The Single-loop Direction Finder for Aircraft. *a. Description.*—

Radio direction finders utilizing the conventional rotating loop designed to make them adaptable to aircraft have been developed and are in use on Coast Guard planes. These outfits, described in a preceding section (Sec. 301c) in this chapter, require a trained operator for their operation and are not adaptable to general use by itinerant pilots. On the other hand, aircraft flying regular airways use radio-beacon receivers

which are not directional in themselves. Until the development of the radio direction finder to be described in this section, the itinerant pilot had no radio aid in his navigation.

The features of this device, which is known as the *single-loop-antenna* radio direction finder to differentiate it from various double-loop-

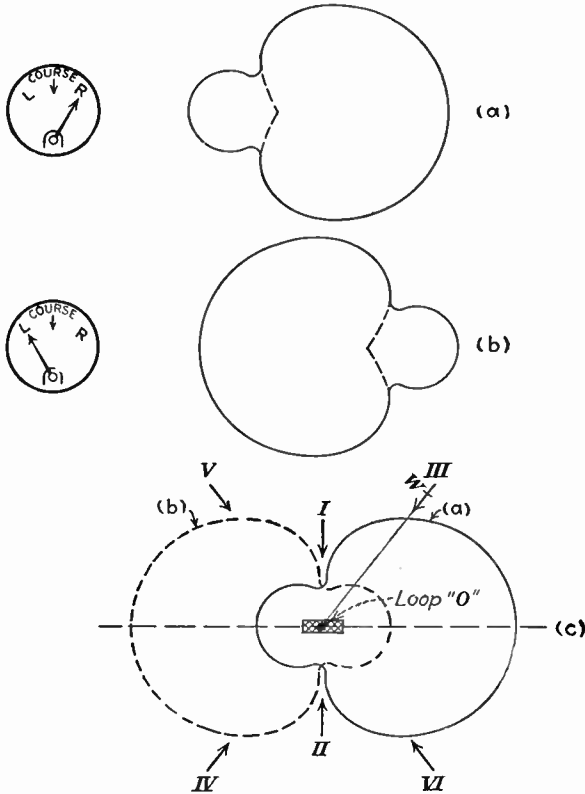


FIG. 336.—Characteristics of unsymmetrical loop.

antenna radio direction finders previously used with unsatisfactory results, may be enumerated as follows:

1. It gives semiautomatic accurate bilateral indication.
2. It can be added to the standard radio receiver of the aircraft.
3. It gives definite directive courses with no ambiguity.
4. It operates on both modulated and unmodulated radio waves from any source (thus broadcasting stations can be used as beacons).

b. Principle of Operation.—A loop which is not symmetrical to the ground acts partly like a vertical antenna in addition to exhibiting inherent loop characteristics. The result is a modified cardioid characteristic as shown in Fig. 336(a) and (b) the dotted lines showing the perfect cardioid without the antenna effect present, the total effect being

produced by the phase difference between the loop and antenna-effect currents. Assume further that this characteristic is secured by unbalancing the loop by grounding one end of it, which we shall call end *A*, the ungrounded end being called *B*. It is logical, therefore, that, if now end *B* is grounded instead of end *A*, a characteristic opposite to that shown in *a* will result. This is shown as curve *b* where again the dotted line shows the perfect cardioid. If a zero-center meter is caused to deflect to the right when the end *A* of the loop is grounded producing curve *a*, and to the left when end *B* is grounded producing curve *b*, and, if, furthermore, ends *A* and *B* are grounded alternately in rapid succession, it is obvious that a bilateral indication of the direction of the received signals is available on the meter. It is the function of the direction-finder circuit to provide for this alternate grounding of the loop ends, and the way in which this takes place, will be explained in connection with the theory of operation of the entire circuit which is shown in Fig. 337.

c. Theoretical Operation of the Direction-finder Unit.—Let us first analyze the operation of the rectifier unit during the period in which no incoming radio-frequency signals are being received. An alternating e.m.f. of an audio frequency is developed by the oscillator circuit and the tube V_3 . These low-frequency oscillations in the primary winding of the transformer T_1 induce an alternating e.m.f. across the secondary winding S of this audio-frequency transformer. During the period in which the e.m.f. across S is such that a positive potential is applied to the plate of the tube V_1 and a negative potential to the plate of the tube V_2 , a current will flow from the cathode of V_1 to the plate, through the radio-frequency choke RFC_1 through the winding S to ground, through the secondary winding of the audio-frequency output transformer T_2 , through the portion X of the resistance R_1 , returning to the cathode to complete the circuit. This flow is indicated by the heavy arrows. When the audio-frequency cycle across S of T_1 changes, the positive potential which operates tube V_2 reverses, and the current flows in the direction of the dotted arrows. During each half-cycle therefore, a current passes through the resistance R_1 , first in one direction and then in the other as the cycle changes. This also results in a current flow through the meter M , first in one direction and then in the other, since it is connected in parallel with the resistance R_1 . However, since the audio-frequency cycle across the resistance and the meter is continually reversing, the current through the meter will be balanced out and the resultant reading will be zero.

Now, when a radio-frequency signal e.m.f. is developed across the tuned circuit L_1C_1 , these voltages are applied to the grid and cathode terminals G and K of the radio-frequency amplifier tube V_4 which results in radio-frequency variations in its plate circuit and through the primary winding of the radio receiver. These radio-frequency signal variations are then amplified through the regular receiver in the general manner, with the exception that, instead of feeding the output signal into an audible indicating device such as a head telephone receiver, the signals are fed back into the special direction-finder unit through the medium of an output transformer T_2 . This results in amplified signal voltages being developed across the secondary winding S of T_2 . Now, with this signal present across this winding, let us see what the effect will be when both the audio-oscillator and the signal voltages are combined.

When one of the rectifier tubes V_1 or V_2 receives a positive potential from the audio-frequency oscillator its plate-to-cathode conductivity is quite high, which results in effectively grounding one end of the loop antenna through the secondary

of T_1 . In other words, the loop is effectively grounded at the end in which the positive potential is applied to the rectifier-tube plate. Similarly, when the audio-frequency potential reverses to the other tube, then the loop is effectively grounded at the other end. Two field patterns are thus alternately produced at an audio frequency as illustrated in Fig. 336(c). Owing to the antenna effect introduced by the rectifiers, these field patterns take the form of a modified cardioid characteristic of a broad or periodic nature.

Thus, when the tube V_1 is functioning, the current passing through the loop produces one field pattern while, when the other tube V_2 is functioning, the other

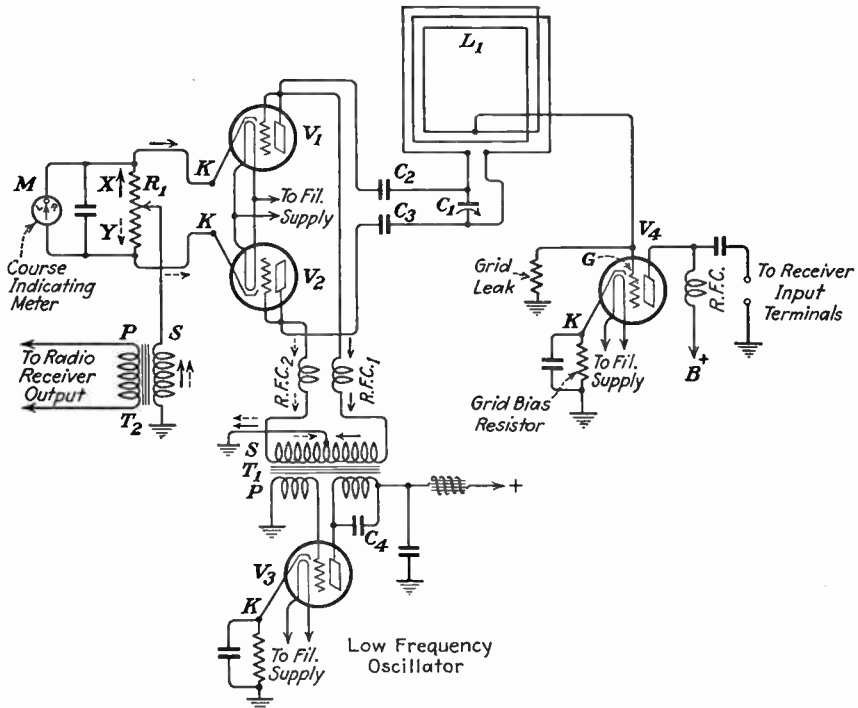


FIG. 337.—Circuit diagram of single-loop direction finder for aircraft.

pattern is produced, forming a typical figure-of-eight characteristic. Assume V_1 is operating, then, when the incoming-signal voltages are induced across the loop and amplified through the regular receiver and across the secondary winding of T_2 , the signal will be proportional to the field pattern at that moment, owing to the fact that the loop antenna is effectively grounded at one end. Consequently, the current passing through the meter indicator M will be proportional to the combined e.m.fs. of the oscillator and signal voltages. Now, when the rectifier cycle reverses and V_2 operates, then the other end of the loop antenna will be effectively grounded and the current through the indicator M will pass in the opposite direction. Since the cycle is continually reversing at the frequency of the audio oscillator, the two e.m.fs. are alternately applied to the indicator M . It may thus be readily seen that the currents tending to deflect the indicator to the right or left must be directly proportional to the difference between the signals due to each pattern. If the pointer of the meter, therefore, is fixed at the center of the scale, the heading of the craft to the right or left

may be readily determined. A variation of 10 degrees from the line of flight will produce practically a full-scale deflection to the right or left with an average adjustment of volume on the receiver.

d. Sense Effect.—Signals arriving from points I and II in Fig. 336(c) give an "on-course" indication on the meter. Now, if a signal comes from point III, the current through the meter will be the sum of the current produced by the curves *a* and *b* or *OW*. Also, as *OW* lies to the right of the on-course line I-II, the meter will deflect to the right, and the loop must be rotated toward the right for an on-course indication. Likewise, if the signal arrives from the direction *V*, the meter will read to the left of the on-course reading. Therefore, when heading toward the source of signal, the meter deflection indicates the direction the loop must be turned for an on-course reading, and a directive sense effect is secured.

e. Construction and Performance.—This direction finder is made up as a complete unit which may be placed near the standard radio receiver to which it is attached. It has a single-control knob which varies the loop-tuning condenser C_1 . The standard radio receiver must also be tuned to the desired signal, but this is not difficult as all other signals are absent owing to the preselection characteristics of the radio-frequency amplifier in the direction-finder unit.

During operation a constant tone of low intensity is present when flying on course. If off-course indications are received, this tone becomes louder thus giving an audible check on the course indication.

The small loop used with this unit is located outside the fuselage immediately over the instrument location or as close thereto as possible.

Some error is introduced by aircraft structure, but this is usually very slight and the accuracy of this direction finder was found on test to be of the order of 1 degree.

Signal intensity is controlled by the volume control in the standard receiver.

The original models of this direction finder utilized a fixed loop. Hence, this compass was used mainly as a "homing" device. More recent models employ a loop which may be rotated, thus permitting the taking of bearings on stations not in the direct course line.

CHAPTER XIII

RECTIFIER UNITS

If a high-voltage direct current could be taken directly from the power lines leading to a radio transmitter or receiver, a rectifier and its associated apparatus would be unnecessary. Unfortunately high-voltage direct current is not universally available for the operation of vacuum tubes, and, excepting for low-powered battery-operated and motor-generator-powered installations, rectified alternating current must be used.

Such a conversion of power requires a rectifier, which is a device having an asymmetrical-conduction characteristic, and which is used for the conversion of alternating current into a pulsating direct current. It is the purpose of this chapter to explain the theory and operation of rectifier apparatus as used to supply power to modern radio equipment.

The parts of a rectifier system in the order of their functioning are (a) the power transformer, (b) the rectifier, (c) the filter, and (d) the voltage divider (if used).

I. THE POWER TRANSFORMER

312. General Transformer Considerations.—The principles of the transformer have been treated in Sec. 42 and will not, therefore, be further discussed here. There are, nevertheless, certain points which apply to a rectifier transformer that need explaining.

a. Heating in a transformer used for rectifier service is more pronounced for a given power than when the same transformer is used as a conventional converter of alternating current. This is due to the rectangular shape of the pulses flowing in the secondary which have a greater heating effect than a pure sine wave. While it is possible to design a transformer which will run cool, it is much more economical, and not disadvantageous, to allow it to run warm. The heat, of course, must never be allowed to rise to the point where the insulation of the transformer would be damaged. Very high power transformers are usually oil cooled. When oil cooling is unnecessary, air cooling is used on all types, the cases being corrugated to provide a greater heat-radiating surface.

b. Direct-current Saturation.—If an alternating current is passed through a transformer winding, the flux will change position with each alternation. Likewise, if an iron core has two coils on it, with the current

in the coils flowing in opposite directions and the current flow taking place in each coil alternately, a similar action of flux reversal will take place owing to the reversing magnetic effect of the coils. On the other hand, if pulses of direct current are caused to flow through a transformer winding, the first few pulses will function to arrange the flux in the magnetic path in a certain direction, and the succeeding pulses will have less and less work to do in this respect because there is no reversing force to undo the work of each pulse. This condition may exist in a rectifier transformer.

As a transformer depends for its operation on the number of lines of force or flux which ebb and flow around the magnetic path, any reduction in this flux movement will reduce the effect of the transformer. It is apparent, therefore, that, if current pulses flow through a transformer winding in one direction only, direct-current saturation will take place eventually. In other words, if direct-current pulses are caused to flow in a transformer winding, eventually the flux in the magnetic path of the transformer will become more or less fixed in one position, and the pulsating current will have but little effect in causing the flux to move.

Direct-current saturation is considered a loss and is likely to take place to a greater or less degree in any transformer winding in which the current pulses flow always in the same direction. Such a condition of current flow takes place in the single-phase half-wave rectifier circuit (Fig. 341), and in the three-phase half-wave rectifiers with star-connected secondaries (Fig. 351).

To prevent direct-current saturation, the magnetic path of a transformer or choke coil may be interrupted by an air gap of a few hundredths of an inch. This is necessary in transformers only when they are used in circuits where there is a possibility of direct-current saturation. As there is always the possibility of such saturation in a ripple-filter choke (Sec. 332), owing to the direct current flowing through it, these chokes are usually designed with an air gap in the magnetic path of the core material to stabilize the inductance value within a given range.

Another method of preventing direct-current saturation in high-power rectifier transformers is to wind the secondary in sections or pies and then interlace the sections; that is, every other section is connected in series to form one side of the center tap.

c. The number of windings and the particular construction of each depend on the service to which the transformer is to be put. A straight power transformer for use in stepping up or stepping down the voltage may require but a single primary and a single secondary circuit. Such a transformer could be used in a half-wave rectifier as in Fig. 341. Usually, if the transformer is to be used in a full-wave rectifier circuit, the secondary must be center-tapped as is T_1 , Fig. 342.

In low-power rectifiers, such as are used in receivers, it is the usual practice to combine the plate and filament transformers into one unit. One or more filament windings may be necessary depending on requirements. It may be necessary to provide a separate winding for the radio-frequency tubes, another for the audio-frequency tubes and still another for the rectifier filaments. All of these secondary windings operate from a single primary winding.

In high-power rectifiers intended for use with transmitters, and where a considerably higher voltage is used on the plate circuits, it is necessary to operate the filaments from a separate transformer as can be verified by studying some of the transmitter diagrams given in this book. It is

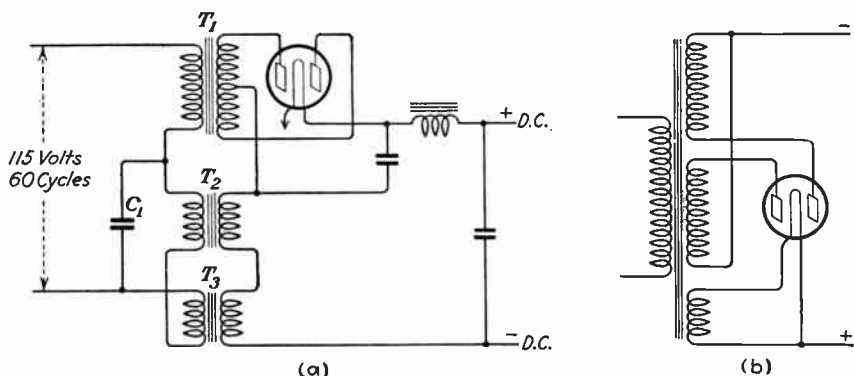


FIG. 338.—(a) Voltage-regulating transformer circuit; (b) reversed-winding method of shielding.

important that each winding be able to carry the load imposed on it in the particular circuit in which it is used. In choosing or designing a rectifier transformer all of these points must be considered.

d. Rating of Transformers.—The rating data of a power transformer are contained on its nameplate and usually specify the input voltage and frequency, and the output volts, current, and power. It is common practice to rate general power transformers as to their safe output in kilovolt-amperes. The reason for this is explained in connection with the section on Kilovolt-amperes (see Sec. 70).

313. Voltage-regulating Transformers.—The theory of operation of the voltage-regulating transformer centers around the principle of the reduction of impedance in a reactor when its core is saturated. A schematic circuit diagram of the scheme employed is shown in Fig. 338(a). In this particular transformer there are two small transformers T_2 and T_3 which supply the regulation properties of the unit. As shown in the diagram these two small transformers have the secondary winding of each placed in series with the negative return of the rectifier circuit. Their two primary windings are placed in series with the primary of the high-voltage transformer.

In the ordinary high-voltage transformer not supplied with regulating windings the voltage delivered by the secondary will tend to drop in value as the current drawn by the load is increased. In the voltage-regulating transformer circuit here shown a complex action takes place which maintains the voltage delivered by the secondary at a steady value regardless of the load within certain limits. This regulating action may be explained as follows: When the current drawn by the load is increased, the flux in the cores of regulating transformers T_2 and T_3 is increased toward saturation; this reduces the reactance of the primary windings of these transformers permitting a higher voltage to be applied to the primary of the high-voltage transformer, thereby maintaining the value of the secondary output voltage of this transformer constant.

It is necessary to use the two secondary windings of T_2 and T_3 in opposition, and they must be very closely balanced owing to the fact that the alternating current across the primaries produce an alternating current in the two secondaries which are in series with the direct-current supply portion of the rectifier, and if they do not buck out, a marked ripple is applied to direct-current output.

The condenser C_1 is placed across the two units for the purpose of shifting the phase relations, and by proper selection of the capacity of this condenser the regulation of the transformer can be varied.

314. Electrostatically Shielded Transformers.—The function of the electrostatic shield which is usually placed between the secondaries and the primary of some high-grade radio transformers is to prevent the transfer of any high-frequency disturbances in the power line from reaching the tube circuits connected to the secondary and also to prevent any high-frequency disturbances in the tube circuits from being transferred back to the line. This shielding is grounded to the transformer casing which in turn is connected to the ground, thus effectively by-passing all electrostatic e.m.f.s. passing between these two circuits to the ground. Therefore, the electrostatic shield in a transformer tends to reduce key-click surges, tunable hums, and to keep radio-frequency currents from entering the power circuits.

It is also possible to secure some shielding action in power transformers used with radio receivers by winding one of the filament windings having a grounded center tap between the primary and secondary windings.

Another means of shielding is to place the outside windings of the secondary at ground potential by means of the *reversed-winding* method in which the two halves of the high-voltage winding have their *outside* terminals connected to the ground instead of to the plates of the rectifier which is the usual procedure. The two *inside* connections are connected to the rectifier plates as illustrated in the diagram, Fig. 338(b).

315. Protective Devices.—It is considered good practice to place fuses in the primary circuit of a power transformer to protect it against

excessive overloads. These fuses are generally rated about 50 per cent in excess of the normal load current.

Surge gaps and arcing horns are sometimes fitted to the transformer secondary to provide a path for high overload currents which might otherwise break down the secondary insulation.

II. THE RECTIFIER

316. The Function of the Rectifier.—As its name implies, the purpose of the rectifier is to rectify alternating current into direct current, pulsating in character but nevertheless flowing in only one direction. A rectifier functions in this way because of its ability to allow current to flow more easily in one direction than in the other.

Although there are many types of rectifying devices operating on mechanical, electrolytic, thermal, and thermionic or electronic principles, the radio operator or engineer is mostly concerned with the so-called *thermionic-type* rectifier utilizing vacuum tubes as the rectifying element. It is to this type, therefore, that the discussion here will be principally, but not entirely, confined.

The theoretical operation of the two-element tube as a rectifier is as follows:

A fundamental rectifier circuit is shown in Fig. 341. As soon as the filament F is heated and emits electrons, these electrons are drawn over to the plate P when the plate is positive with respect to the filament, resulting in a current flow through the circuit. If this current path is traced out, it is seen that, starting at the lower end of the transformer secondary, the current flows to the lower terminal of the output, through the load to the upper terminal of the output, thence to the filament and across to the plate by electronic conduction back to the upper end of the transformer completing the circuit.

It is important that we get polarity relations straight in our minds at this point. We know that polarity is a measure of potential between two points, and that current flows from negative to positive. Therefore, if we make an instantaneous observation of polarity during the half of the alternating-current cycle when the rectifier tube is passing current, we would find (a) that the plate is positive with respect to the filament, (b) the upper end of the transformer is positive with respect to the lower end, and (c) the upper output terminal of the output is positive with respect to the lower terminal. Considering the output, it is now clear why the positive terminal of the rectifier is taken from the filament or cathode of the rectifier and not from the plate. This becomes clear from studying the diagram.

On the next alternation of alternating current the potential in the secondary of the transformer is reversed, and the plate of the tube becomes negative with respect to the filament. Under these conditions the tube does not conduct current, instead it (the tube) takes the form of an infinite resistance. Thus only those alternations flowing in the direction in which the tube conducts are allowed to pass, and pulsating direct current results.

A graph showing the effect of the rectifier on the alternating current is shown in Fig. 339(b). The alternating current entering the transformer is shown as a sinu-

soidal wave form (a); the rectified direct current delivered to the output is shown as a pulsating wave form (b). This is the wave form at the output before the filter is applied (disregard the pulses marked I and II for the present, they will be referred to in Sec. 318).

317. The high-vacuum rectifier tube is a two-element tube, essentially the same as the conventional vacuum tube in construction but with the grid omitted. Into this tube a high vacuum is introduced, and all possible traces of gas are removed to prevent the tube from breaking down under high-potential strains.

The two most important characteristics in connection with this rectifier are: (a) maximum peak plate current, and (b) maximum peak inverse voltage.

The maximum peak plate current is the safe value of current that the tube can handle under continuous operating conditions. Normally it is

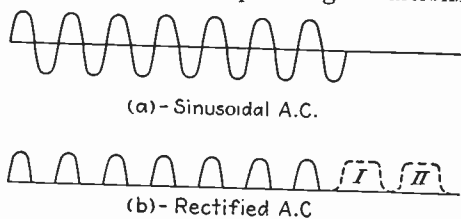


FIG. 339.—Effect of half-wave rectifier on alternating current.

governed by the cathode emission permissible with a given tube. As the direct current flows in a rectifier tube for only half the time the tube is in operation, this current as indicated by the plate milliammeter should never be allowed to exceed more than half the maximum peak plate-current rating of the tube.

The maximum peak inverse voltage is the potential applied across the tube while it is non-conducting. In other words, it is the voltage across the tube during the alternation the tube is idle. For example, in Fig. 339 the tube is idle during each negative alternation (below the line) of sinusoidal alternating current; hence no rectified alternating-current pulses result during these periods. As the study of rectifier circuits is made, it will become clear that this maximum peak inverse voltage is practically the full voltage of the transformer secondary in single-phase circuits.

An important factor in connection with the high-vacuum tube is the voltage drop across the tube during operation. When the cathode is heated, electrons are drawn toward the plate and a space-charge voltage drop is built up in the tube which depends upon the current drawn and the construction of the tube (see Sec. 73). This voltage drop may vary from a few volts to several thousand volts. Any circuit calculations involving this tube must take this variable drop into consideration. Furthermore, the regulation in a circuit using this tube is poor and is governed largely by the drop in the tube.

The high-vacuum rectifier tube utilizes a tungsten filament and requires a relatively high filament power. A few types of high-vacuum rectifier tubes utilizing an oxide-covered filament have been developed for low-power use, the most common of these being the half-wave type UX-281 and the full-wave type UX-280.

In high-power rectifiers of the high-vacuum type, it is necessary to dissipate the heat generated in the tube by a water-cooling system which increases the cost and the space required.

The characteristics of various sizes of high-vacuum rectifier tubes are shown in Table XII in Appendix A.

318. The Mercury-vapor Rectifier Tube. *a. Description.*—One of the greatest American contributions to modern radio power-supply development has been the perfection of the low-pressure mercury-vapor rectifier tube. In order to better understand the reason for this some of the disadvantages of the high-vacuum type of rectifier tube will be stated, and the way the mercury-vapor tube has eliminated these disadvantages will be pointed out.

One of the chief obstacles to the efficient use of the high-vacuum-type tube is the variable-drop characteristic which causes the regulation of the tube to be very poor. This is an especial handicap in circuits where the load varies over a wide range as in class B amplifiers. The high tube drop also generates heat which must be dissipated in the larger size tubes by a water-cooling system. The filaments, being made of tungsten, require a large power.

The mercury-vapor tube has solved many of these difficulties. This tube under normal operating conditions has a constant tube drop of 15 volts, regardless of load conditions. Because of this low tube drop there is no excessive heating of the tube, and even the large models of the mercury-vapor tubes are air cooled. At the same time, and for the same reason, the efficiency of the mercury-vapor tube is very much higher. In one comparison made, the efficiency of the high-vacuum-type tube rectifier was 87.5 per cent, while that of the mercury-vapor tube rectifier was 98.8 per cent.

The cost of the mercury-vapor installation is also much less than the cost of a high-vacuum-type tube installation partly because the water-cooling system is not necessary on high powers.

The two most important characteristics of the mercury-vapor tube are:

1. The maximum peak inverse voltage at which the tube can be operated without flashing back.
2. The maximum peak plate current at which the tube can be operated without filament disintegration.

Since the voltage drop in the tube is only 15 volts it can be neglected, and the inverse voltage is practically the full potential of the transformer secondary in a single-phase circuit. The peak plate current depends

on the circuit conditions. In a single-phase circuit the peak current, when the rectifier is connected to a resistance load, is simply the maximum of the sine wave of the rectified pulses. If inductance is placed in series with the load, the output-current curve takes the form of successive square blocks, as shown by the pulses I and II of curves *b* in Fig. 339 and the peak plate current is nearly equal to the output current. If a condenser is connected across the rectifier output, the peak plate current may reach values equal to from three to five times the output current, and for this reason a condenser is rarely connected across the output of a mercury-vapor rectifier tube, which is sometimes designated by a black arrow head opposite the tube symbol on the diagram.

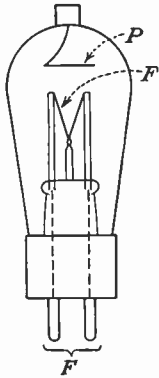


FIG. 340.—
Mercury-vapor
rectifier tube,
low-power type.

b. Theoretical Operation.—In the high-vacuum-type rectifier tube the electrons which are emitted by the cathode must overcome a space charge, which surrounds the cathode like a shield, before they can reach the plate (see Fig. 77(a)). This space charge is the cause of the large voltage drop between the elements and is the principal disadvantage of the high-vacuum-type tube. In the mercury-vapor tube, mercury-vapor gas at low pressure is introduced into the tube. The presence of this gas changes entirely the theoretical operation of the tube. A view of the construction of a low-power tube (type RCA-866) is shown in Fig. 340. The filament- and plate-voltage supply connections are shown in Fig. 341. The elements are placed so as to reduce the possibility of flashover, the plate being located at the top of the tube with its connection being made to a cap.

When the filament of the tube is lighted, the mercury vapor is raised to operating temperature. The high-voltage plate circuit may now be closed and the tube will operate. When the electrons are emitted by the filament and are drawn toward the plate, they collide with atoms of mercury vapor. The collision of an electron with an atom of mercury vapor knocks one of the atom's electrons outside of the influence of its nucleus, and the atom, being minus an electron, becomes a positive ion. This positive ion is immediately drawn over to the space charge surrounding the filament and neutralizes a negative ion in the space charge. In the normal operation of the mercury-vapor tube there are enough positive ions released from the mercury vapor practically to neutralize the negative space charge surrounding the filament. In fact it reduces the space charge drop to 15 volts, and it remains at this value, regardless of load, during the normal operation of the tubes.

From a practical standpoint the 15-volt drop holds as true for the largest size tube as for the smallest. This means that the voltage regulation of the rectifier unit depends only on the regulation of the transformer, supply, and circuit.

As long as a positive potential is applied to the plate, current will flow through the tube from filament to plate. Therefore, if an alternating potential is applied to the tube, rectification will result. When the plate is negatively charged, it repels what free electrons are in the vicinity as the filament is now positive and the space charge neutralized; and as the cold plate does not emit electrons there can be no ionization of the mercury vapor owing to collision and no flow of current through the tube.

c. Practical Operation.—Great care must be taken not to apply the positive potential to the plate before the filament has warmed up sufficiently to build up a space charge, as ions attracted to the filament under these conditions might strike the filament with such a velocity as to dislodge active material from the filament and

possibly ruin the filament entirely. For this reason rectifier circuits employing the mercury-vapor tube are fitted with a time-delay relay which delays the application of the plate potential until the filament has had time to reach incandescence and the proper space charge is formed.

The mercury-vapor rectifier tube cannot be substituted for the high-vacuum-type tube unless the rectifier apparatus, filters, and associated equipment are capable of withstanding the higher output voltage resulting from the lower tube drop. For example, if the terminal voltage at the transformer terminals is 700 volts and the tube drop when using a high-vacuum-type tube is 50 volts, under certain load conditions, the output of the rectifier will be 650 volts. Now, if a mercury-vapor tube with a drop of only 15 volts is used, the output of the rectifier will be 685 volts. The equipment must be able to withstand this higher potential if a substitution of tubes is to be made.

The operating characteristics of mercury-vapor tubes are shown in Table XII in Appendix A.

As the filament of the tube is called on to emit electrons to the full extent of the plate current, the filament or cathode must be ruggedly made. In the mercury-vapor tube the filament is coated with oxides of barium and strontium, and it is the evaporation of this oxide coating that determines the life of the tube. High-power tubes are given a heavier coating of oxide than are the tubes designed for lighter loads. If the tube is abused by application of the plate potential before the filament has been brought to operating temperature, the heavy ionic bombardment of the filament which takes place under these circumstances may ruin the filament in a very short time by disintegrating its oxide coating quickly.

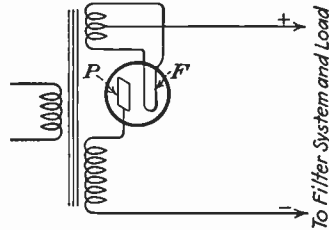


FIG. 341.—Single-phase half-wave rectifier.

319. Inverse-peak Voltage.—An examination of the full-wave rectifier circuit shown in Fig. 342 will show that when tube 1 is conducting, the full secondary voltage, minus the drop in tube 1, is applied to tube 2. As this is during the non-conducting alternation as far as tube 2 is concerned, it is the inverse of the alternation during which the tube operates as a conductor of current. For this reason it is called the *inverse-peak voltage* applied to the tube to differentiate it from the voltage applied when the tube operates to pass current.

Every rectifier tube has a rating as to the safe inverse-peak voltage at which the tube may be operated. If the maximum peak inverse voltage of a high-vacuum rectifier tube is exceeded, the plate of the tube will get unduly red but no permanent damage will result. If, on the other hand, the maximum peak inverse voltage of a mercury-vapor tube is exceeded, the potential difference between the plate and the filament will exceed the safe limit (usually about 22 volts) and disintegration of the filament will occur, permanently damaging the tube. It is evident, therefore, that, when the mercury-vapor tube is used, great care must be taken in designing the circuit so that the safe inverse voltage will not be exceeded.

Let us return again to a consideration of the circuit. If the mercury-vapor type of tube is used, the drop in the tube is 15 volts. There-

fore, if, for example, the output of the transformer secondary is 5,000 volts, the inverse-peak voltage applied across the non-conducting tube is 4,985 volts, or very nearly the full secondary voltage. Let us now assume a high-vacuum rectifier is used instead of the mercury-vapor tube. The high-vacuum tube may have a drop of from only a few volts to several thousand volts depending on the load conditions. For example, let us say the tube has a drop of 100 volts in the circuit shown in Fig. 342. If the transformer secondary yields 5,000 volts the maximum peak inverse voltage across the non-conducting tube is 4,900 volts, the 100 volts being lost in the tube passing the current.

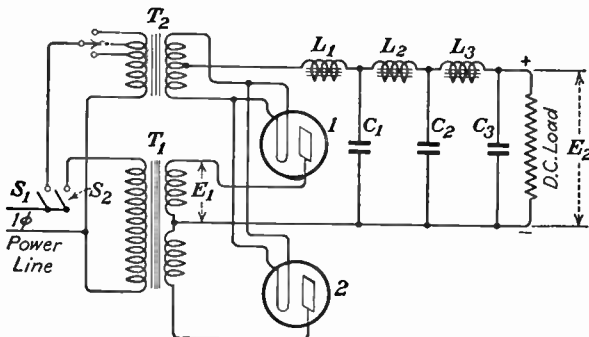


FIG. 342.—Single-phase full-wave rectifier.

Notice further that the inverse voltage applied to the non-conducting tube is the *full* secondary voltage. In other words, if the voltage each side of the center tap is 2,500 volts, then the *full* voltage is 5,000 volts. This will become clear from a study of the diagram. Notice that the plate of tube 1, for example, is connected to one end of the transformer secondary. Now when tube 2 is conducting, the filament side of tube 1 is connected to the other side of the transformer secondary (the lower end in the diagram, Fig. 342) through tube 2 which is now almost as good a conductor as a closed switch, its drop being only 15 volts. Also, if the full voltage of the secondary is 5,000 volts, the secondary of the filament transformer must be insulated to withstand this potential strain as it is the positive side of the rectifier output and is above ground potential.

320. Rectifier Circuits Classified.—Rectifier circuits are divided into several classes, namely:

- | | | |
|--------------|---|--------------------------------|
| Single-phase | { | (a) Half-wave circuits. |
| | | (b) Full-wave circuits. |
| | | (c) Bridge circuits. |
| | | (d) Voltage-doubling circuits. |

In addition to these single-phase circuits the following three-phase circuits are in use:

Three-phase $\left\{ \begin{array}{l} (a) \text{ Half wave.} \\ (b) \text{ Half-wave double } Y. \\ (c) \text{ Full wave.} \end{array} \right.$

321. Single-phase Rectifier Circuits.—The operation of the single-phase half-wave rectifier has already been described in connection with Fig. 341 which shows the complete half-wave circuit including the filament connections. As the filament of the tube becomes the positive side of the rectifier output, it is essential that the filament-transformer secondary be insulated to withstand the full peak voltage of the rectifier system.

A *full-wave single-phase circuit* is shown in Fig. 342. As soon as the filaments of both tubes are heated by the filament winding, an electronic

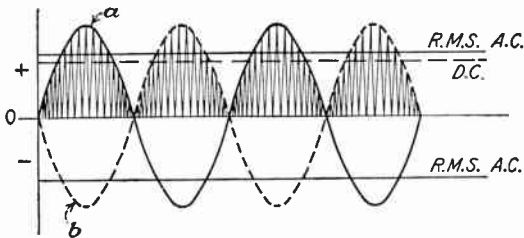


Fig. 343.—Effect of full-wave rectifier on alternating current.

emission from both filaments results. However, during the period that the top side of the secondary winding is positive, the bottom side will be negative. Hence, during this period, electrons are only attracted by the plate of the top tube, and a current flows through tube 1 from filament to plate. Tube 2 is idle. Then, as soon as the alternating-current e.m.f. reverses its polarity, the lower end of the high-potential transformer becomes positive (+) and the top end negative (-). The current then flows through tube 2, and tube 1 is idle.

Thus, if we consider the polarity of the secondary winding as constantly reversing at a definite frequency, the two rectifier tubes function alternately; that is, when one tube receives a positive potential on the plate, the other tube is negative on the plate, and, as the polarity changes, the condition is reversed. Hence, the current flowing through the load is practically a constant direct current.

In the half-wave rectifier system it will be recalled that the rectified wave shape was of an intermittent character [see Fig. 339(b)], that is, the peaks did not follow in rapid succession. However, in the full-wave system the peaks are actually doubled, due to the filling-in action of the second tube. This may readily be seen by referring to the wave shape of a full-wave rectifier system illustrated in Fig. 343.

In this set of curves (Fig. 343), the potential on tubes 1 and 2 is represented by the sine curves *a* and *b*, respectively. The r.m.s. secondary voltage applied to each tube is represented by the lines *R.M.S.* This is the alternating-current voltage indicated by a meter across each half of the secondary, reading from the zero line in the curve diagram in each instance. The shaded portion represents the rectified pulses of output direct current, available as output voltage. The dotted line *D.C.* is the output voltage of the rectifier. This is the value indicated by a direct-current voltmeter across the output and is 85 per cent of the r.m.s. value of secondary voltage delivered to the tube in this circuit.

Figure 344 illustrates the mean current effect in a circuit using either pure direct-current, half-wave, or full-wave, rectification. Here it will be seen that the current flow in a full-wave rectifier system more closely resembles the pure direct-current flow such as is obtained with a B battery.

System (a) produces a rather heavy current fluctuation.

System (b) produces a considerably lower degree of current fluctuation.

System (c) produces practically no current variation as long as the plate-supply batteries are in good condition.

From these illustrations it will be seen that the current fluctuations, even with the full-wave rectifier, are still too great to insure satisfactory operation when used as a plate supply for vacuum-tube receivers. It will, therefore, be necessary to smooth the variations or ripples to a still greater degree. This is accomplished by the use of a filtering system consisting of inductances and capacities later to be described (Sec. 328).

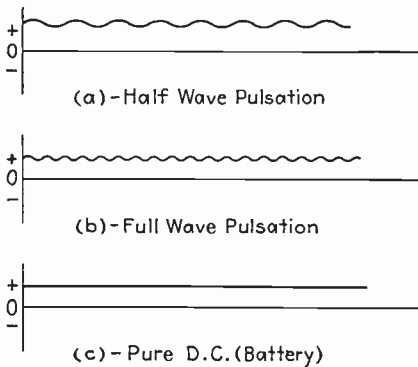


FIG. 344.—Comparison of rectified and battery e.m.fs.

322. Bridge-rectifier Circuit.—

The bridge circuit is shown in Fig. 345. It is seen that this is a single-phase full-wave rectifier which requires four tubes. A simplification

of this circuit is shown in Fig. 346. Referring to either figure, the path of the current during one alternation is through tube 4, the load, tube 1 to the other side of the line. During the next alternation the current flows through tube 3, the load, tube 2 to the other side of the line. This circuit is used when a high-output potential is required because the strain is equally divided between the two tubes passing the current; the inverse voltage is likewise equally divided between two tubes materially reducing the strain on each.

Compared with the two-tube full-wave rectifier, this four-tube circuit utilizes the complete secondary voltage during the whole of each cycle and, therefore, gives twice the output voltage available from the

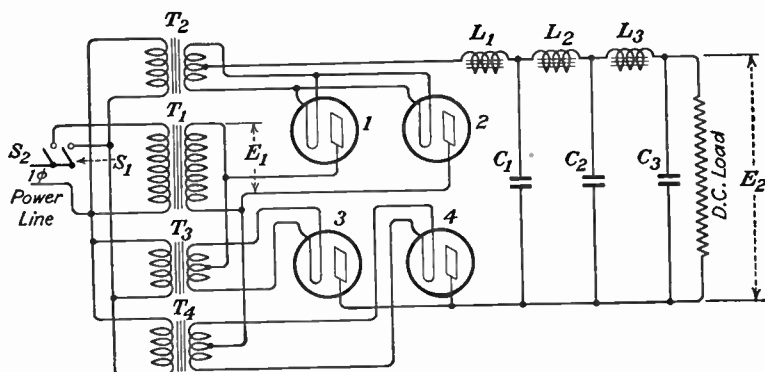


FIG. 345.—Four-tube single-phase full-wave rectifier, bridge type.

same transformer when only two tubes are used each utilizing one-half the transformer secondary through the center tap.

323. Voltage-doubler Rectifier.—The two-tube half-wave voltage-doubler rectifier is shown in Fig. 347 and is similar in some respects to the bridge circuit as will be seen by referring to the diagrams in Fig. 348(a). Here it is seen that two condensers C_1 and C_2 replace the tubes used in the conventional bridge circuit. A study of this circuit will show that these condensers are charged alternately, first one and then the other, and are discharged in series thereby doubling the output voltage. The output voltage depends entirely upon the size of

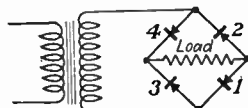


FIG. 346.—Equivalent circuit of bridge-type rectifier.

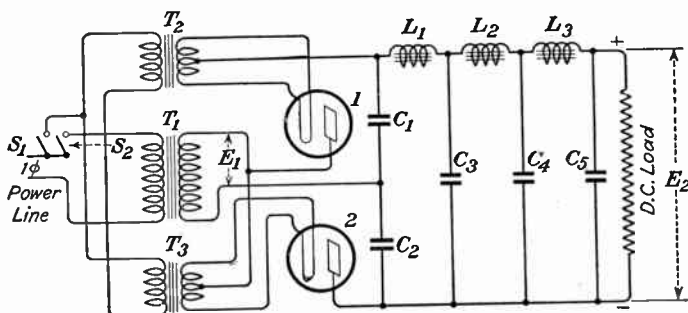


FIG. 347.—Two-tube half-wave voltage-doubler rectifier.

condensers used and the current drawn by the load.¹ The maximum current drawn should not exceed the rated capacity of one tube.

¹ For method of calculation of voltage-doubling circuit see Armstrong, Polyphase Rectification Special Connections, *Proc. I.R.E.*, January, 1931, p. 93.

A voltage-doubler circuit using four tubes is shown in Fig. 349. The total voltage rating in this circuit is twice the rating of a single tube. A graph of the output-voltage wave form of the voltage-doubler circuit is shown in Fig. 348(b).

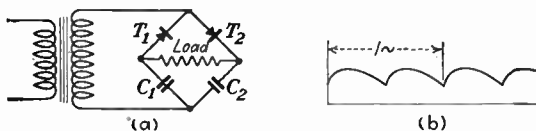


FIG. 348.—Equivalent-circuit voltage-doubler rectifier.

This circuit is useful where an inexpensive rectifier is desired to produce high-voltage, low-current outputs. It may be used with either high-vacuum-type tubes or with mercury-vapor tubes. If it is used with mercury-vapor tubes, care must be taken that the inverse-peak voltage does not exceed the safe limit of the tube.

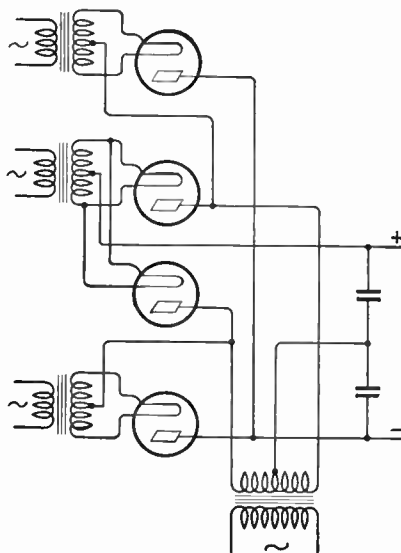


FIG. 349.—Four-tube full-wave voltage-doubler rectifier.

324. Three-phase Rectifiers. a.

Introduction.—In Sec. 71 on Polyphase Alternating-current Circuits it was pointed out that there are two methods of connecting apparatus to a three-phase circuit. These are the delta and star methods, the latter being also known as the Y or wye connection. It is suggested that Sec. 71 be reviewed now in order to bring out clearly the voltage and current relations in both the delta and star connections. *The reason for the factor 1.73 in the computation of current and voltage relations in a three-phase circuit must be clearly understood.* This method of computation applies also to three-phase transformer circuits.

There are four basic methods of connecting three-phase transformers as shown in Fig. 350. The effect of these systems on the value of the rectified voltage is as follows: The *delta-delta* and the *star-star* connections give only the step-up effect of the transformers and may be expressed $E_s = E_p \times T_r$, where E_s is the secondary voltage, E_p is the primary or line voltage, and T_r is the voltage transformation ratio of the transformer.

In the *star-delta* connection the output voltage is expressed

$$E_s = \frac{E_p \times T_r}{1.73} \quad (174)$$

The reason for this is easily seen; the primary of the transformer is star connected which means that the line voltage is not connected directly across any single primary winding but is rather applied to two windings in series. This decreases the voltage in each phase winding to the value

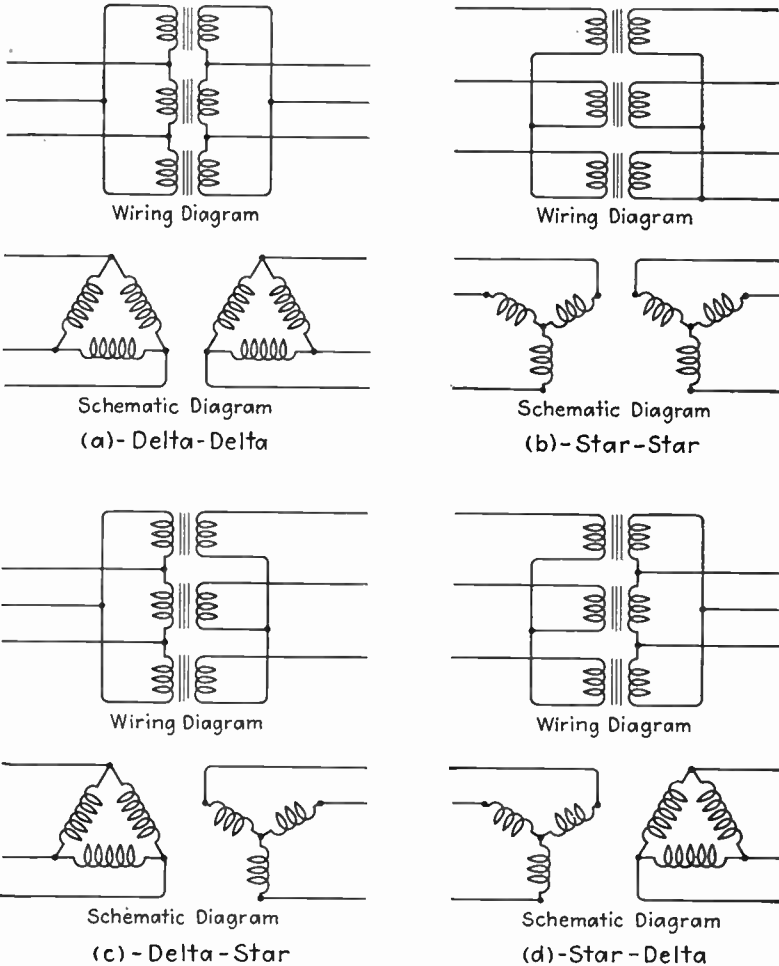


FIG. 350.—Transformer connection methods.

of the line voltage divided by 1.73. The secondary windings, being delta connected, delivers the full secondary voltage to each phase line. Therefore, it is seen that the output voltage is only 1/1.73 of the voltage obtained with either the delta-delta or the star-star connection.

In the *delta-star* connection the output voltage is expressed

$$E_s = E_p \times T_r \times 1.73 \tag{175}$$

Here we have the opposite effect to that produced by the star-delta connection. The effect of the primary on the line voltage is nil because each primary coil is connected across a phase of the line and receives full voltage value from it. The secondary, being star connected, does not receive the full effect of induction in each coil from each primary coil, because it (the secondary) is connected two coils in series across each output line. The voltage across each line, owing to two secondary coils in series, while it is not twice the voltage produced by one secondary coil, is very nearly that. To be exact, the voltage is increased by the factor 1.73 in addition to the step-up voltage factor of the transformer.

b. *The half-wave three-phase rectifier* is shown in Fig. 351. It is to be noticed that, in this instance, the delta-star transformer connection is

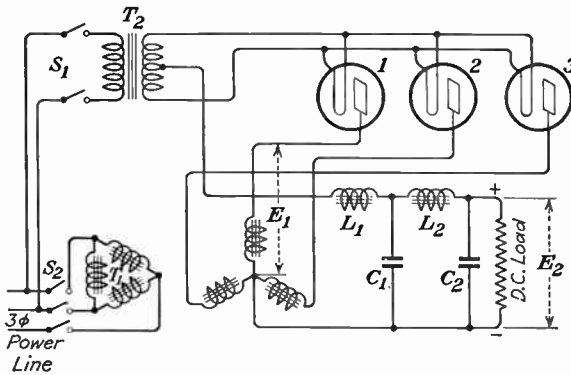


FIG. 351.—Three-phase half-wave rectifier.

used in order to take advantage of the extra voltage step-up secured by this method. Each tube carries current one-third of the time, and the effect on the output current is to give three pulses of direct current for each three-phase cycle of alternating current. The direct-current output is limited by the maximum safe plate-current rating of the rectifier tubes used and is computed by formula (175).

c. *The half-wave double-Y three-phase rectifier* is shown in Fig. 352. This circuit is in effect the same as two three-phase half-wave circuits operating from a single three-phase primary winding with the two sets of secondary windings connected so that the voltage is delivered to the plates of the two sets, of three tubes each, 180 degrees out of phase. The current is equally divided between the two sets of tubes by the *inter-phase reactor* (L_1) connected between the centers of the transformer secondaries.

In this circuit advantage is also taken of the delta-star method of transformer connection to give increased voltage output. As the tubes are operated in parallel, this circuit is capable of delivering twice the current of the three-tube circuit. Each tube works on one-third of the

cycle, but, as the tubes connected to each secondary network are active 180 degrees out of phase, the output wave form is equal to that of a full-wave three-phase rectifier. This circuit does not, however, have the voltage-doubling characteristic of the full-wave rectifier which will be explained next.

d. The full-wave three-phase rectifier is shown in Fig. 353. This is the latest of the three-phase circuits to be developed and is a considerable improvement over the half-wave circuits shown before. Here again the

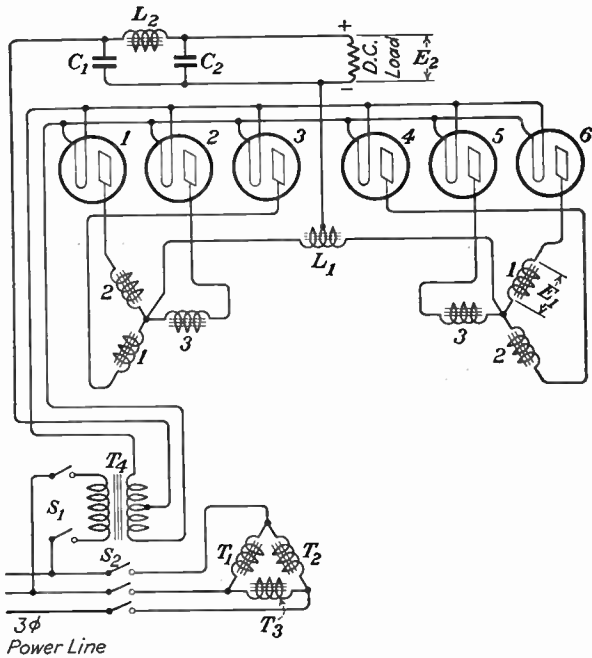


FIG. 352.—Three-phase half-wave double-Y rectifier (equivalent to six-phase output).

delta-star method of connection is employed to get maximum voltage increase. This circuit is so efficient that the total effect of the rectifier is to more than double the voltage over and above the step-up effect of the transformers. This circuit is in effect a voltage doubler. This may be explained as follows. Owing to the efficiency of the three-phase, full-wave circuit, $95\frac{1}{2}$ per cent of the transformer secondary-voltage peaks¹ is available in the output of the rectifier. Also, as the delta-star system of transformer connection is employed, the voltage output is again increased by the factor 1.73. Therefore, the output voltage of the rectifier may be calculated

$$E_s = E_p \times T_r \times 0.955 \times 1.4 \times 1.73$$

¹ Peak voltage in this circuit is taken to equal 1.4 times the r.m.s. value of secondary voltage across each phase.

or

$$E_o = E_p \times T_r \times 2.31 \tag{176}$$

which gives the rectifier more than a voltage-doubling characteristic.

A principal advantage of this rectifier circuit is that the maximum peak inverse voltage applied to the tubes is only $4\frac{1}{4}$ per cent greater than

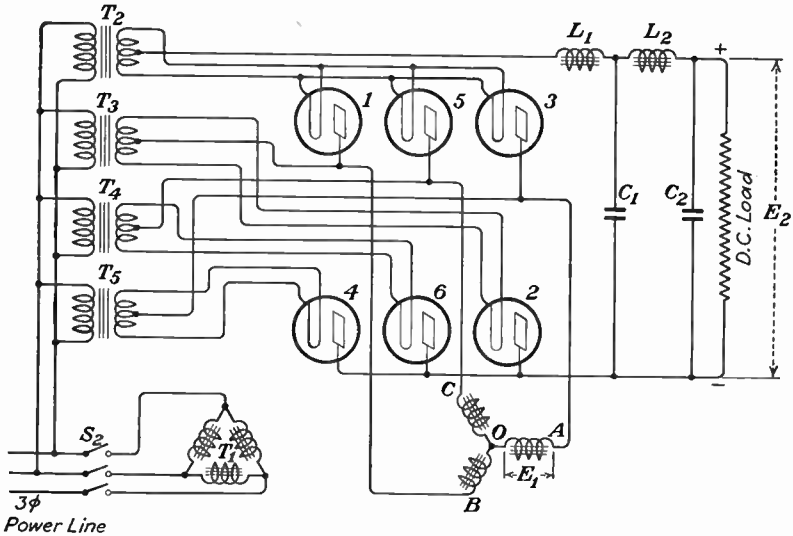


FIG. 353.—Three-phase full-wave rectifier.

the average output voltage. The output wave form of this rectifier is a six-phase ripple which is easy to filter into an almost pure direct current.

e. Output Voltages of Three-phase Rectifiers.—A composite set of curves showing the output-voltage forms of three-phase rectifiers is shown

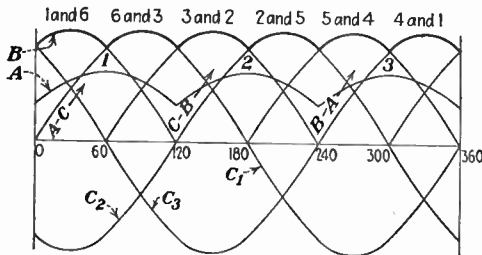


FIG. 354.—Output voltages of various rectifiers compared.

in Fig. 354. Curve A shows the output from the half-wave circuit of Fig. 351. The characteristics of this circuit are such that only about 48 per cent of the maximum peak voltage is available as average output voltage. All these curves show the voltage conditions with a resistive load and no filter.

Curve *B* shows the output voltage from the half-wave, double-Y rectifier of Fig. 352 and from the full-wave circuit shown in Fig. 353. Both of these circuits give a six-phase ripple which does not require a large filter in order to give a pure direct-current output. The order in which the tubes function is indicated by the numbers at the peaks of each pulsation. The letters *A-C*, *C-B* and *B-A* on the secondary e.m.f. curves (Fig. 354) indicate the path of the current through the transformer secondary on the full-wave rectifier. The labels *C*₁, *C*₂, and *C*₃ indicate the first, second, and third cycles, respectively.

325. Mercury-arc Rectifiers.—A diagrammatic illustration of this type of rectifier is shown in Fig. 355. The glass bulb or tube forms an

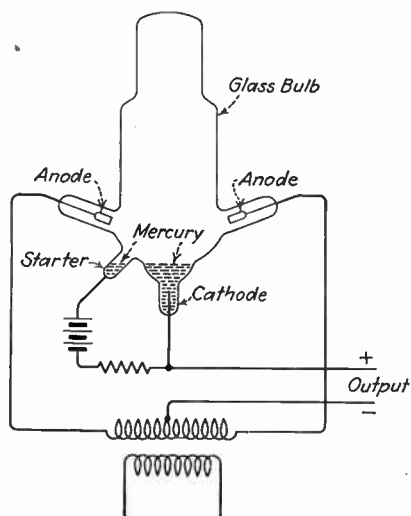


FIG. 355.—Mercury-arc rectifier tube.

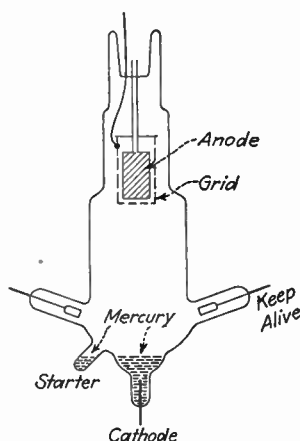


FIG. 356.—Mercury-pool cathode-power grid-glow rectifier tube.

enclosing envelope for the anode terminals, the cathode, which is a pool of mercury, and the starting electrode, which also terminates in a pool of mercury. To start the tube operating, it is necessary to tilt it so that the two pools of mercury make electrical contact. This completes the "keep-alive" circuit which receives energy from the battery connected in the circuit. When the tube is brought back to the vertical position, an arc forms as the mercury breaks contact, which causes an ionization of mercury within the tube and provides a path for the high-potential current supplied by the transformer to flow from the cathode to the conducting anode via the load. As the rectifier tube shown is a full-wave type, one anode acts on each alternation. Likewise one-half of the secondary operates on each alternation, both the anodes and secondary sections operating alternately, current flowing only to the positive anode from the secondary section to which it is connected.

Small-size mercury-arc rectifiers are glass enclosed, while large sizes are metal enclosed. The smaller sizes of mercury-arc rectifiers have been largely replaced by the mercury-vapor tube because it is not necessary to provide tilting apparatus to start the latter. Large-size heavy-duty metal-enclosed mercury-arc rectifiers are in present-day use to supply power to high-power radio transmitting stations. These metal-clad rectifiers incorporate the use of a grid for voltage control.

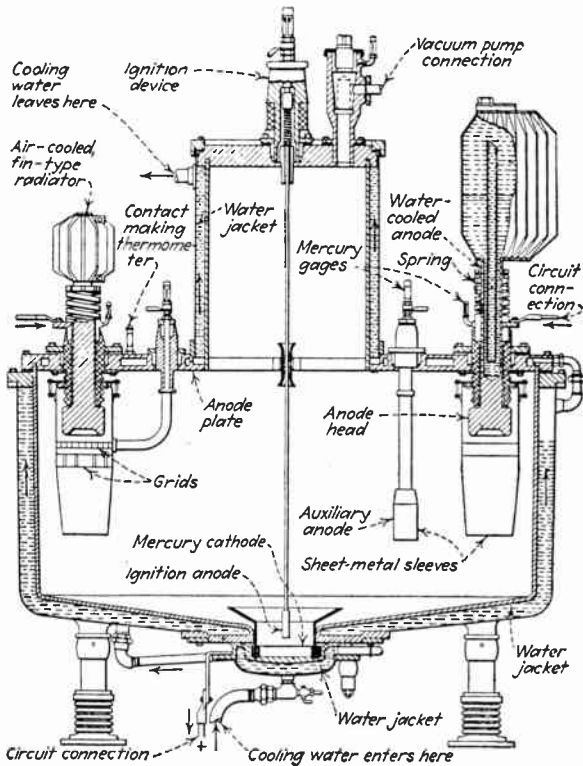


Fig. 357.—Metal-clad mercury-arc rectifier. (Courtesy of Allis-Chalmers Manufacturing Company.)

Grid control of voltage has also been incorporated in glass-enclosed mercury-arc rectifiers. These tubes are known as *mercury-pool cathode-power grid-glow tubes*. Such a tube is shown in Fig. 356.

326. Metal-clad Mercury-arc Rectifiers.¹—The rectifier here described is of *Brown Boveri* design and is used to convert alternating-current power into high-voltage direct-current power. It is grid-controlled and capable of standing heavy temporary overloads without damage.

¹ Manufactured by the Allis-Chalmers Manufacturing Co., Milwaukee, Wis. This information is condensed from manufacturer's engineering leaflet 2167. Readers desiring full information on these equipments should send for this leaflet.

a. *Description.*—A metal-clad mercury-arc rectifier consists essentially of an iron cylindrical tank in which a vacuum is maintained by a mercury-vapor pump in conjunction with a rotary oil pump. Figure 357 shows a cross section through a grid-controlled rectifier of this type. The various parts are plainly designated. The tank is divided into an arc chamber and a condensing chamber. At the base of the arc chamber is a cathode consisting of a pool of mercury. Mounted on the cover plate of this chamber by insulator bushings are the main anodes, two small excitation anodes, and grid elements associated with each main anode. The condensing chamber is located above the arc chamber and is closed at the top by a cover plate carrying an ignition device and the vacuum-pump pipe-line connecting flange.

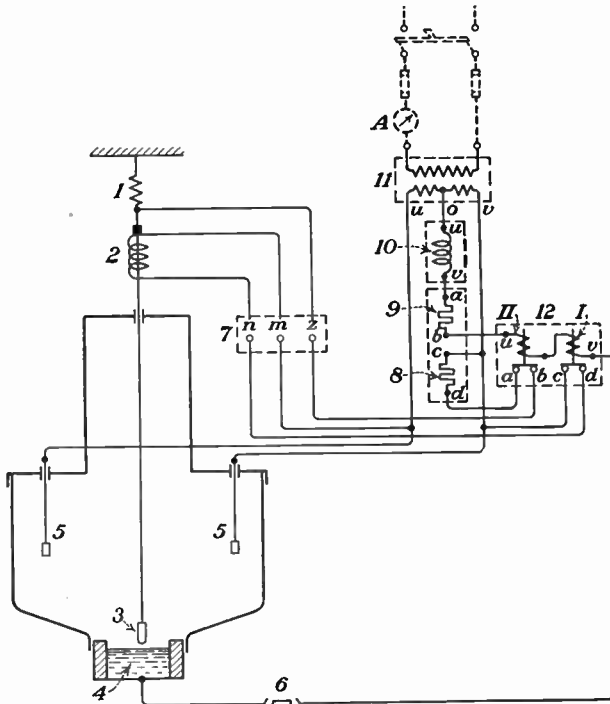


FIG. 358.—Connection diagram metal-clad mercury-arc rectifier. (Courtesy of Allis-Chalmers Manufacturing Company.)

Two types of main anodes are shown in the cross-section diagram. Water-cooled anodes are employed only on units supplying high-amperage direct-current power, whereas anodes equipped with air-cooled fin-type heat radiators are employed on high-voltage units.

Rectification depends upon the characteristic valve action of a mercury arc, this valve action allowing current to flow in one direction only. Each anode of a multi-anode rectifier carries the arc current during the time it is at highest positive voltage in relation to the other anodes. The mercury vaporized by the arc is condensed and due to the force of gravity automatically returns to the cathode.

b. *Operation.*—When a rectifier is first assembled in the workshop, it is pumped to a high vacuum and thoroughly “degassed.” It is then sealed for shipment. Before being put into service, a second degassing process is employed which usually takes only a few hours time.

In starting up a mercury-arc rectifier, an arc must first be struck by the ignition anode. This is accomplished in less than 1 sec. by supplying alternating current to a solenoid which by magnetic force causes the excitation anode to dip momentarily into the mercury pool of the cathode. The ignition arc struck at the instant of withdrawal of the ignition anode is automatically transferred to two excitation anodes supplied with low-voltage alternating-current power, after which two relays remove power from the ignition apparatus. Figure 358 shows a diagram of connections of the ignition and excitation circuits. The low-voltage excitation arc is maintained independently of the main arc during operation to prevent the necessity of having to reignite the rectifier in case the main load current momentarily drops to zero. This is a particularly necessary feature in a rectifier supplying high-voltage direct-current power to the transmitting tubes of a radio-telegraph station in which frequent intervals occur when no power is demanded of the rectifier. The excitation arc is equivalent to the tube filaments in a hot-cathode tube-type rectifier, with the exception that its life is unlimited and that it ordinarily requires only a fraction of the power demanded by the filaments of an equivalent tube-type rectifier.

When a metal-clad rectifier is shut down for a period of only an hour or two, it may be restarted immediately at full power. After longer shutdown periods, it is generally advisable to start a metal-clad rectifier at reduced voltage or to bring it up to nearly normal operating temperature by means of a small low-voltage heating transformer provided for the purpose. This requires about 10 min. time.

Figure 359 shows the essential parts of a circuit diagram of a metal-clad grid-controlled mercury-arc rectifier plant as employed in a radio transmitting station to furnish high-voltage direct-current power to the plate circuits of the transmitting tubes. In the upper part of the diagram, a typical radio-transmitter circuit is shown in elementary form, and in the lower part the rectifier circuit with associated power, control, and auxiliary equipment is illustrated.

327. Junction Rectifiers. *a. Description.*—Rectifiers of this class depend for their operation on the unilateral conductivity of certain metals when brought into contact. Two of these rectifiers which are finding practical use are the copper oxide rectifier and the magnesium-copper sulphide rectifier. The current flow in both types is from the mother copper to the copper compound, which results in the positive output terminal being connected to the copper and the negative terminal to the copper compound. The metals must make contact under high pressure in order that the separation between them shall not be more than the distance comparable to the molecular separation in the two metals.

b. Copper Oxide Rectifier.—The most recently developed type of copper oxide rectifier is based on the physical fact that, if pure copper is brought into atomic contact with cuprous oxide, the unit thus formed will offer a much higher resistance to the flow of an electric current in one direction than in the other. Actually this resistance is on the order of 3 to 1, that is, three times the resistance is offered in one direction than in the other. Rectification takes place between the copper and the cuprous oxide, and this action is electronic and not electrolytic in nature.

In order to work efficiently the copper oxide rectifier must operate cool, and for this reason the individual rectifier junction is made up in the form of a washer about $1\frac{1}{2}$ in. in diameter to which may be fitted cooling

flanges if the rectifier is to be used for relatively high currents. Each washer is made of copper which is oxidized on one side. When assembled,

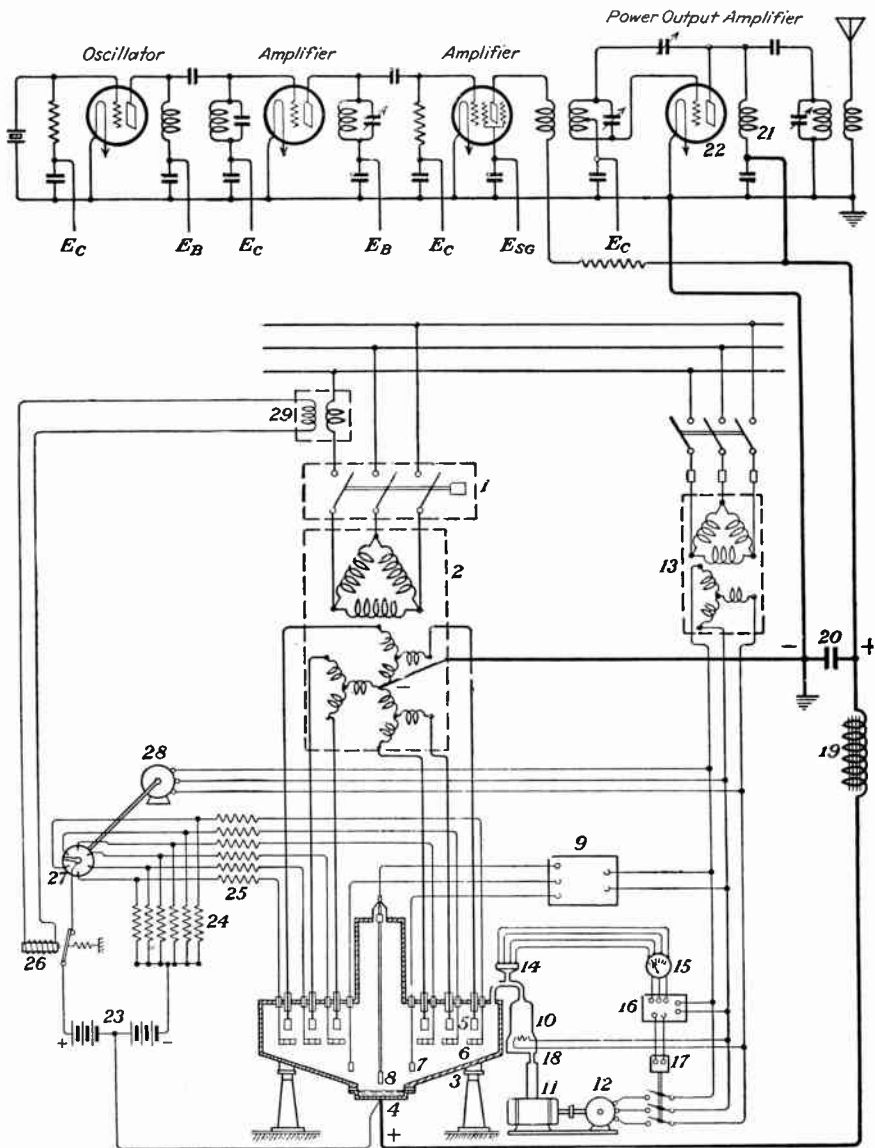
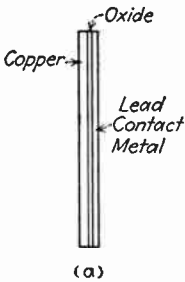


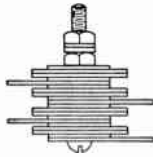
FIG. 359.—Power-circuit diagram utilizing metal-clad mercury-arc rectifier. (Courtesy of Allis-Chalmers Manufacturing Company.)

electrical connection is made to the oxide by a washer of lead or metal foil. The construction and assembly of the rectifier are shown in Fig. 360.

Under practical conditions this type of rectifier works at efficiencies of 60 to 70 per cent. Although it is possible to assemble any number of units to give any practical power rating, this type of rectifier finds its greatest usefulness where only low voltages and power are required. The breakdown voltage per junction is 11 volts alternating current, and the breakdown temperature is about 160°F. The life of the rectifier will



(a)



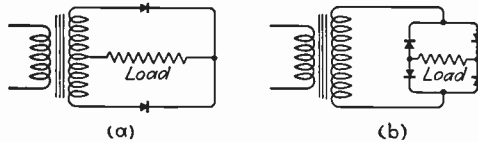
(b)

FIG. 360.—Junction rectifier unit.

run into the thousands of hours, if the operating temperature is kept below 140°F. Tests have shown a life of 6 years for this rectifier with little or no deterioration. For proper operation the pressure on the units must be correct, and for this reason the rectifiers are assembled at the factory and held together by a rigid clamp construction. The pressure should

not be altered after the unit leaves the factory.

The copper oxide rectifier is made up commercially for use in meter conversion where it is used when it is desired to convert a direct-current meter into a universal meter for measuring both direct and alternating current. In this manner any direct-current milliammeter can, with the addition of this rectifier unit and the proper resistances, measure alternating- as well as direct-current voltages and current. It is also used together with other accessories such as a transformer and condensers to replace the dry cell in supplying current to the microphone, and for this purpose units are made to supply 15 ma. per button.



(a)

(b)

FIG. 361.—Two ways of connecting copper oxide rectifiers.

This unit is also used to supply direct current to excite the field coils of dynamic speakers. It was formerly used to a great extent for battery charging and for B eliminators but the tube-type rectifier has largely replaced it for these purposes.

In broadcast-station operation the copper oxide rectifier is finding use as a low-voltage filament rectifier for station amplifiers, condenser microphones, and voltage amplifiers. It can be made to deliver six or twelve volts at current drains up to seven amperes.

The two standard ways of connecting the copper oxide rectifier are shown in Fig. 361. The bridge circuit in diagram (b) is perhaps the more desirable circuit for this type of rectifier as it reduces the strain on each unit by one-half.

c. The magnesium-copper sulphide rectifier is a dry junction-type rectifier which is made up of disks of magnesium and cupric sulphide in the form of washers assembled on a bolt and clamped together. A rectification ratio of 75:1 giving a resistance of 30 ohms to current in one direction and a resistance of 0.4 ohm in the other direction is secured. Proper operation depends on the elements being clamped together with a pressure of about 200 lb., and this pressure should not be changed by the user.

The safe operating voltage per junction is 4 volts, and the units are designed to operate at 90°C. (about 190°F.). The operating temperature may be raised to 150°C. without injury to the unit. The units are self-healing for minor breakdowns and the life runs about 5,000 hr. There is, however, deterioration if this rectifier is left idle in a moist atmosphere.

This type of rectifier is best adapted to bridge circuits. Like the copper oxide rectifier before described, this rectifier has been largely replaced by tubes. It is finding some use, however, in dynamic loudspeakers for field excitation, and a more recent development has been the adaptation of this rectifier to voltage doublers and quadruplers for experimental and miscellaneous uses.

III. THE RIPPLE FILTER

328. Function of Ripple Filter.—The current delivered by a rectifier consists of a direct- and an alternating-current component. For all practical radio purposes this alternating-current component causes too much of a variation in the output of the rectifier to make the power thus delivered fit for use as plate current in vacuum-tube circuits. This variation would cause audio modulation of the emitted wave of the transmitter or of the output of the receiver fed by such a plate supply. It is necessary, therefore, that this current be filtered, that is, its variations smoothed out.

A study of Fig. 362 will make clear the conditions in a typical rectifier circuit as a whole from input to output. Curve (a) is the alternating current flowing in the transformer primary and takes the form of a pure sine wave. In most instances this is a 60-cycle current, although in some places 50 cycles or even 25 cycles is supplied. The curve shown represents by its amplitude a 110-volt potential.

Curve (b) shows the output of the secondary of the transformer, which is also a pure sine wave. The voltage has now been raised to 330 volts. This is the reading that an alternating-current meter would register if connected across the secondary, and is the effective or root-mean-square (r.m.s.) value of the voltage. The peak value of this voltage, that is, the potential strain to which the rectifier apparatus would be subjected, is found by the equation

$$E_m = E_{eff} \times 1.414$$

This formula, when dealing with tube rectifiers, is only of theoretical interest because the output of such a rectifier is erratic in nature and is not a pure sine wave.

To arrive at the true peak values of a tube-rectifier output an oscillographic study must be made of the wave form. When this is not possible a generous safety margin must be allowed in figuring peak voltages.

Curve (c) shows the effect of a full-wave rectifier circuit on the transformer voltage shown in the preceding curve. Notice that there are two pulses of direct current for each alternation of alternating current. It is just as if the sine wave

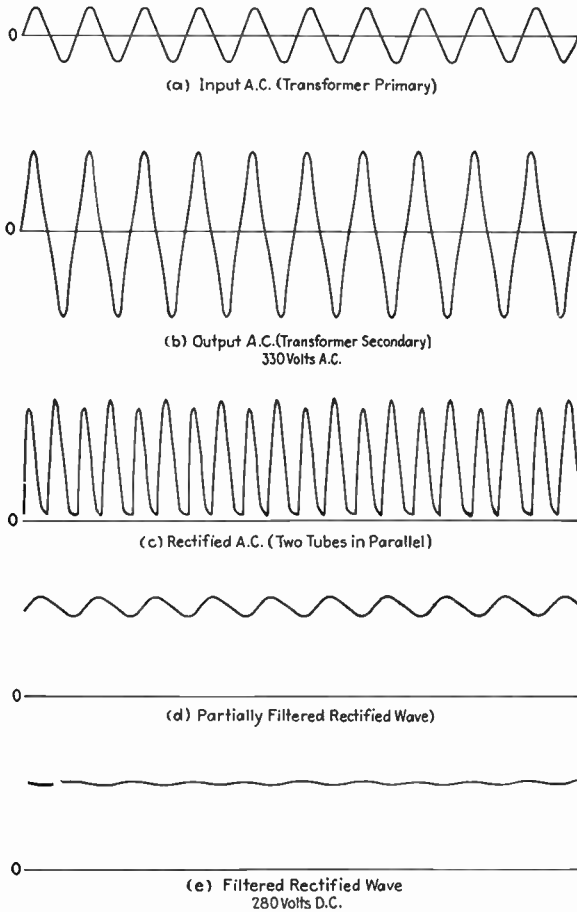


FIG. 362.—Rectified alternating current from input to output.

of the secondary voltage were folded along the zero line, the negative alternations becoming positive pulses by the rectifier action. It is interesting to note, too, that the effect of one tube is not exactly similar to the effect of the other tube, which accounts for the unevenness in the peak values of the rectifier because the tubes are not exactly balanced. In the particular curve shown the unbalance is not bad, but it serves to illustrate another source of unevenness in the output of the rectifier. Each cyclic variation tends to set up another alternating-current component which must be filtered by the filter.

Before discussing curves (d) and (e), the filter will be considered in detail.

All filters used in rectifier circuits are low-pass filters and as such are designed to pass currents of all frequencies below a critical or cut-off frequency and to reduce substantially the amplitude of currents of all frequencies above this critical frequency. This is shown graphically in Fig. 363, which is a graph of the typical performance of low-pass filters such as are shown in Fig. 364. The curve shows that when the particular low-pass filter represented by this graph is inserted in the line, the insertion loss at 20 cycles is 10 db, at 50 cycles, about 20 db, at 60 cycles, 24 db at 120 cycles 35 db, and so on. Hence, it is seen that any low-frequency hum which gets through the filter is of a very low power value, the exact level depending, of course, upon the initial value of the ripple frequency.

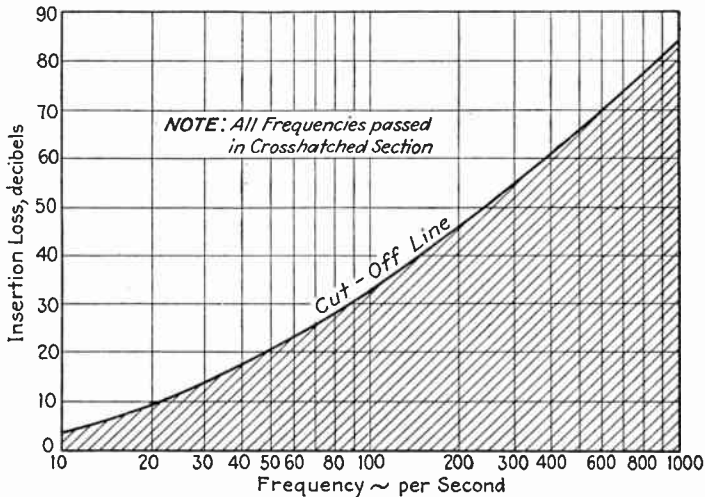


FIG. 363.—Low-pass filter performance graph.

This curve also shows the desirability of a high-frequency ripple output in that the high-frequency ripple suffers greater attenuation from the filter and, therefore, is easier to eliminate. The note on the graph reads, "All frequencies passed in cross-hatched section." A consideration of the situation will show this to be true. For example, suppose the ripple frequency is that from a three-phase full-wave rectifier such as is shown in Fig. 353, namely, a six-phase ripple resulting in 360 cycles. By referring to the curve it is seen that the 360-cycle frequency is passed by the filter; but the loss is equal to 58 db. The initial power intensity would therefore have to be tremendous in order to be present in the output at all, and such a condition is unlikely in any situation. The rectifier filter is made up of audio or power chokes and condensers designed to withstand a high-potential strain. This type of filter is also known as a *ripple filter* because it reduces the ripple current, while freely passing the direct current from a rectifier or generator.

The ripple filter consists of a series-arm impedance which is always a choke coil or coils, and a shunt impedance which is always one or more condensers. The series arm, therefore, is always inductive reactance and the shunt arm is capacitive reactance. For method of deriving the values of L , C , and f_c , see Sec. 152.

Various types of ripple-filter circuits are shown in Fig. 364. The input or high end of the filter is designated by the numbers 1 and 2, and the output or low end of the filter by the numbers 3 and 4. Likewise, the line terminating in terminals 1 and 3 is termed the *high side* of the line and the line terminating in the terminals 2 and 4 is called the low or ground side.

A ripple filter may have a condenser input as in Fig. 364 (c) and (e), or it may have a choke-coil input as in circuits (a), (b), and (d). There is

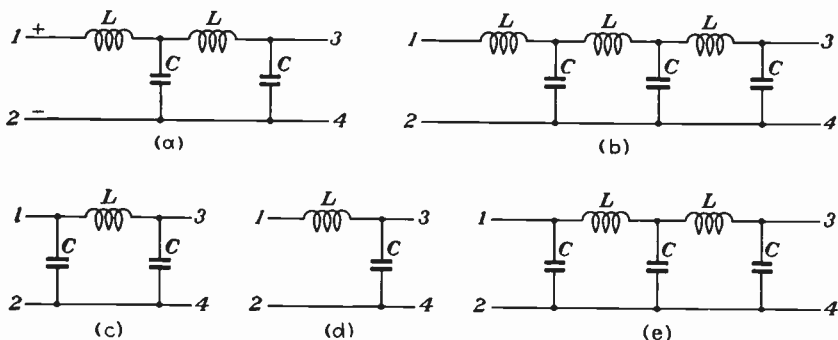


FIG. 364.—Various low-pass filter circuits; [all inductances (L) are iron-core chokes].

a considerable difference in the action of these two types of circuits as will be seen later. The condenser-input filter is used only with the high-vacuum type of tube rectifier while the choke-coil input filter is used with the mercury-vapor rectifier.

The function of the choke coil in the filter is to smooth out the current fluctuations while the function of the condenser is to smooth out the voltage variations. Both working together constitute a ripple filter. In visualizing the operation of a filter, it is necessary to imagine the effect of inductance and capacity on an alternating current. The choke coil tends to oppose any change in the current in the circuit and depends for its operation on the counter e.m.f. it sets up when an alternating current flows through it. The source of the power is inconsequential; it may come from a direct-current generator and be the alternating-current component set up by the commutator ripple or it may be supplied by a tube rectifier; the action of the choke coil in limiting current peaks is the same in both circuits.

329. The action of the condensers in the filter is not the same, however, regardless of the source of voltage. When a condenser is con-

nected to an alternating-current line which is supplied by a generator, the counter e.m.f. supplied by the condenser, once it is charged, will equal the e.m.f. of the line regardless of the size of the condenser. Also, as the condenser can discharge back through the armature windings of the generator, the voltage across the condenser is always the voltage of the line.

When a condenser is connected across a rectifier, another action takes place. This can best be visualized as follows:

The first pulse of current passed by the rectifier puts a charge on the condenser. This charge is but a fraction of the total charge the condenser is capable of holding. As the tube is unilateral in conductivity, the charge can leak off the condenser only by flowing through the load. This it does slowly. Before it has all leaked off, how-

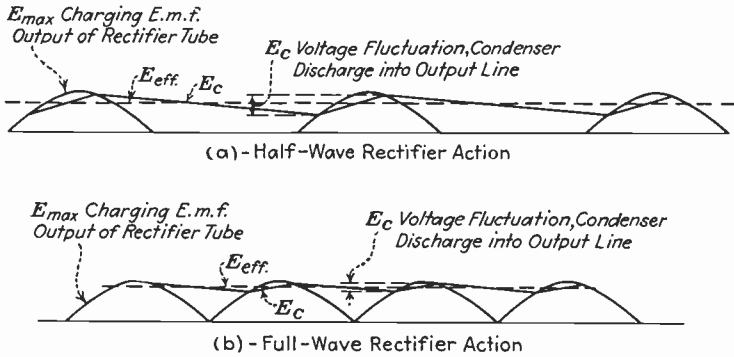


FIG. 365.—Condenser action in filter unit.

ever, another pulse of current comes along and increases the charge on the condenser. Therefore, as the discharge of the condenser is slower than the charge, the condenser finally becomes fully charged up to its rated capacity. Once the condenser is fully charged the discharge action becomes regular as in Fig. 365. As some period of time does elapse between pulses of current, the condenser discharges to a certain extent; so, its potential is lower at the beginning of each pulse than at the end of the pulse. This causes a fluctuation in the value of the voltage fed to the output by the condenser as shown in the curves in Fig. 365, E_c being the voltage fluctuations. The amount of this fluctuation has an effect on the final ripple in the output. Also notice that this fluctuation is greater when the time periods between charging pulses is greater. It varies directly with this factor. For this reason there is less ripple from a full-wave rectifier [curve (b)] than from a half-wave rectifier [curve (a)]. It is also readily seen how the condenser voltage depends on the load. If the load permits the condenser to discharge too rapidly, the rectifier pulses will not charge the condenser rapidly enough to maintain the required voltage. For this reason, the output voltage of a rectifier unit decreases as the load increases and vice versa. To decrease the condenser-voltage fluctuations, the condenser must be large enough in capacity to hold more electricity than can be drained off by the load in the time intervals between pulses. Also the rectifier must be of a sufficient rating so that it will be capable of charging the condenser up to the required potential.

It has been found, from both theoretical solutions and practical tests,¹ that, of the condensers used in condenser-input filters, the first-filter condenser produces the greatest effect on voltage output and regulation but comparatively little effect on the ripple. The greater the capacity of the first condenser, within certain limits, the greater will be the voltage output of the filter circuit, and the better the voltage regulation. The second-filter condenser has a lesser effect on voltage output and regulation but a very important effect on the ripple. The last condenser in the two-section filter shown in Fig. 364 (e) serves primarily as a reservoir to supply momentary high demands of the transmitter.

Working voltage is the term applied to designate the potential at which the condenser should be worked for longest life in the circuit. This is an important factor in the design of ripple filters. A good practical rule is to specify as the working voltage of the first condenser of the condenser-input filter a value three times the output voltage, and for the following condensers a value twice the output voltage of the rectifier.

330. Paper Filter Condensers.—The conventional paper condenser consists of two metallic foils, which constitute the plates of the condenser, separated by a special dielectric paper. These sheets of foil and paper are laid one over the other and wound into a form not unlike the well-known bolt of cloth seen on the shelves of a dry-goods store. This form is then placed in a casing or container depending on whether the condenser is of the ordinary dry-paper type, or whether it is of the oil-impregnated or oil-filled types.

The area of the metallic foils, and the dimensions of the condenser container depend, of course, upon the capacity rating of the condenser. By the use of oil impregnation and filling, the physical dimensions for a given capacity are materially reduced.

In the oil-filled type the condenser is placed in a non-leakable container, and oil is poured in to fill up all air spaces. This oil has special properties, which will be mentioned in a moment, and naturally soaks into the dielectric, thus increasing the capacity of the condenser and its breakdown voltage. In the oil-impregnated condenser, the paper dielectric is impregnated with a special condenser oil having high dielectric strength together with a high dielectric constant similar to that used in the oil-filled type.

The plain paper condenser, not oil impregnated, is satisfactory for receiver power packs, and this type is available with breakdown voltages up to 1,000 volts. The paper condenser has, however, been largely replaced for this purpose by the electrolytic condenser later to be described.

For transmitting rectifying units the oil-filled or oil-impregnated condensers are preferable and are available with breakdown voltages as high as 3,000 volts or higher.

331. Electrolytic Filter Condensers.—The electrolytic filter condenser depends for its operation on the fact that certain metals when used as anodes in certain electrolytes become coated with a very thin insulating film due to polarization. This film acts as a dielectric separating the two electrodes constituting the condenser.

¹ The information in this and the next paragraph is condensed from a report on filter condensers in the *Aerovox Research Worker*, vol. 4, no. 3.

The *wet-type electrolytic condenser* is enclosed in a container which also holds the liquid electrolyte. Into the container is placed a number of corrugated aluminium plates which are bent into ridges to increase their surface area and, consequently, the capacity of the condenser. These aluminium plates are all connected together to form one electrode (anode) of the condenser. The electrolyte forms the other electrode. A flat aluminium plate is placed opposite each corrugated plate to serve as a means of passing current into and from the electrolyte. A film of oil is floated on top of the electrolyte to prevent evaporation.

Various types of electrolytes will operate in the electrolytic condenser; to mention two, there is ammonium citrate as one, and a solution of borax and boric acid in water as another. The latter is perhaps the most commonly used, it being non-combustible, non-poisonous, and non-injurious to clothing. The condenser works best at temperatures ranging from 40 to 105°F. Any heat over 105°F. will greatly shorten the life of the condenser, best operation being obtained at a temperature of about 75°F.

The electrolytic condenser has a unilateral-conduction characteristic, that is, it has an exceedingly high resistance to current flowing in one direction but a very low resistance to currents in the opposite direction. The voltage which may be impressed across the condenser before the film constituting the dielectric breaks down and permits an appreciable leakage current is called the *critical voltage* of the condenser. The voltage which may be impressed on the condenser safely without danger of rupturing the dielectric film is called the *working voltage*. At the present time the standard working voltage of the electrolytic condenser is 500 volts. This, of course, is below the critical voltage.

When a greater working voltage is desired, several condensers may be placed in series. But as this reduces the capacity, it may be necessary to use a series-parallel arrangement to get the required capacity.

Electrolytic condensers of the wet type find use as ripple-filter condensers on rectified low-voltage filament-supply lines. In view of their unidirectional properties these condensers must be connected to the line with the anode of the condenser to the positive side of the line. If the connections are reversed, heavy currents will flow which would ruin the condenser.

Because of the thinness of the dielectric it is possible to get very high capacity values in a small space. For given areas of electrodes, capacities 1,000 times greater may be obtained with the electrolytic condenser, than it is possible to get with a paper condenser. Electrolytic condensers have a poor power-factor (low efficiency) compared with paper condensers, but this is not a detriment of much consequence in ripple-filter circuits.

Great care must be taken to keep impurities out of the condenser. The containers must be clean, and the aluminium especially pure. Impurities of any kind, especially organic, will materially shorten the life of this type of condenser. To keep the film on the anode, it is advisable to keep the condenser continually across a live circuit, and it should be connected across such a circuit for not less than 1 hr. each week at a minimum.

Dry electrolytic condensers operate on the same principle as the wet type just described but differ somewhat in mechanical construction. The electrodes consist of foil which is an alloy of aluminium and which may be wound into a roll form and placed in a tubular container. These electrodes are separated by a gauze, usually of double thickness, which is soaked with the electrolyte.

One well-known type of dry electrolytic condenser utilizes for its electrolyte a solution of boric acid, glycerine, and ammonia, either gaseous or as an ammonia water, in the proportion of 1,000 grams of glycerine, 620 grams of boric acid, and about 50 cc. of 26 per cent ammonia water or gas. This results in a fluid with which the gauze is impregnated. The dry electrolytic condenser is dry only in the sense

that its electrolyte cannot under any circumstances spill out as it is held by the gauze.

In all other respects the theory of operation of the dry electrolytic is identical with that of the wet electrolytic condenser.

332. The action of the choke coil in a ripple filter is deserving of study because of its important function in the circuit.

If a choke coil of a certain rating in henrys is inserted in an alternating-current line, its rating remains constant so long as the frequency of the line remains constant. The inductive action of the choke, as we have seen before, is due to the action of the lines of force cutting the windings of the choke and setting up a counter e.m.f. which tends to prevent any change in the current flowing through it. It has the same effect on an alternating current that a flywheel has on the motion of a steam engine; it tends to keep its variations at a minimum and its operation smooth. A choke coil placed in a direct-current line would affect the circuit only as a resistance; as there are no moving lines of force about the coil, there can be no inductive effect. Likewise, if for any reason the number of lines of force which cut the coil for a given time are reduced, the inductive effect of the coil is reduced.

A choke coil in a ripple filter has flowing through it principally direct current. In addition to this it has an alternating-current component set up by the moving lines of force about the coil, caused by the rise and fall of each pulse of direct current. When the pulse is rising in value, the lines of force are going out from the coil in a given direction, outward from the center of the wire to infinity. Also, when the pulse is decreasing in value, the lines of force are collapsing on the coil. In this way there is a constantly moving field of force cutting the coil windings, first in one direction and then in the other, resulting in the setting up of an alternating current. It is this alternating current which is unwanted in the output of the power supply, and it is this current which the choke tends to damp out because of its inductive effect.

Every choke coil has a saturation point above which an increase in current will affect only slightly the value of the choke. The direct current flowing through the choke coil in the ripple-filter circuit raises the flux to a certain percentage of the saturation value. If this direct current is raised to a point where it increases the flux of the choke well above the saturation point, it is readily seen that the alternating-current component can have no effect because any change in the flux above the saturation point has no effect on the coil as a choke. If the direct current is now reduced to the saturation point, the coil will have a slight inductive value owing to the decrease in flux below saturation by every negative alternation of alternating-current component; but every increase above the saturation point has no effect, and, therefore, the coil has but little inductive value.

The choke to operate efficiently must not have more than a certain amount of direct current flowing through it. This amount is limited to a point which still allows a margin for the alternating-current component to affect the coil. The change in flux in this margin must be great enough to produce the inductive effect of the coil.

A choke coil for a ripple filter is rated, therefore, to have a certain inductive effect with a certain direct current flowing through it. Any increase or decrease in this direct current will change the inductive value from the rated value. It is important, therefore, to choose a choke coil capable of operating efficiently with the direct-current load the circuit must handle.

As the direct-current resistance of the choke causes a loss of voltage, which must be subtracted from the amount delivered by the rectifier, this resistance must be kept as low as possible.

A choke coil for a ripple filter may be rated 15 henrys, 150 ma., 210 ohms, which means that as long as the current through the choke is not raised above 150 ma. the coil will have an inductance value of 15 henrys. The power loss in the choke may be computed by the familiar I^2R formula, and the voltage drop is figured $E = I \times R$.

Iron-core choke-coil inductive values (henrys) are usually measured with 60-cycle alternating current. There is but little difference in the inductive value of a choke, when the frequency is raised to 120 cycles. For values of frequency higher than 120 cycles the inductive value of the choke would have to be remeasured if an accurate rating is desired. For example, if a choke is rated at 30 henrys it means 30 henrys at 60 cycles. It could still be considered 30 henrys at 120 cycles from a practical standpoint. Therefore, if a 30-henry choke is specified for a single-phase full-wave rectifier, where the ripple frequency is 120 cycles, any 30-henry choke could be used. On the other hand, if a 15-henry choke is specified for a three-phase full-wave rectifier it would mean a choke having an inductance of 15 henrys at a ripple frequency of 360 cycles. The ordinary 15-henry choke would be more than 15 henrys at this frequency and would be a larger choke than necessary.

The effect of a varying load on the inductance value of a ripple-filter choke is controlled by an air gap in the magnetic path of the choke. The best designed chokes keep the inductance value at anticipated amounts from minimum to full load. This "swinging" value in a choke becomes an important factor in class B and C amplifier power supplies where the current drawn by the load varies over a wide range. Choke coils designed to operate with varying loads are referred to as *swinging chokes*.

Retard or retardation coil is another name for a ripple-filter choke, although a retard coil may have other functions if used in another circuit. This is the designation generally used by the Western Electric Company in their apparatus descriptions. The filter choke is also known as a *reactor*, which designation is generally used by the RCA in their apparatus descriptions. A reactor may or may not have an iron core depending on its use in the circuit, although all ripple-filter reactors utilize iron cores.

333. The first choke coil in the circuits used with the mercury-vapor tube [Fig. 364(a), (b), and (d)] has a very important function in preventing current peaks from reaching a value high enough to damage the filament of the rectifier tube. Important experimental work¹ has shown that this choke has a certain critical value which may be expressed for 60-cycle circuits by the equation

$$L_o = \frac{R}{1,000} \quad (177)$$

where L_o is the critical value and R the resistance of the load. The optimum value (the value which works best) in a filter circuit is $2L_o$ or twice the critical value.

In addition to protecting the mercury-vapor tube against excessive current peaks, it improves the smoothing action of the filter, greatly improves regulation of voltage, and reduces heating of the secondary winding of the transformer.

¹ See article by Dellenbaugh and Quimby in *QST*, February, 1932.

The choke connected before the first condenser is a necessity in all mercury-vapor rectifier circuits. As this type of rectifier is rapidly replacing the high-vacuum type in modern equipments, a series inductance or first choke usually precedes the first shunt condenser in the rectifier units accompanying these outfits.

334. Equalizing and Bleeder Resistance.—If it is necessary to connect several condensers in series across the output of a power supply, condensers of unlike capacity ratings should not be used, as under these circumstances the voltage distribution across the individual condensers will be uneven, the smallest condenser taking the highest voltage drop. It is better practice to use condensers of equal-capacity rating when a series bank is used.

When a number of condensers are connected together to form a series bank, it is advisable to connect an equalizing resistance across the entire bank to equalize the voltage drop across each condenser unit. When such a resistance is provided, each unit has a tendency to maintain across itself a voltage equal to that across each of the other units. A schematic diagram showing the equalizing resistor in position is shown in Fig. 366. The purpose of the resistance is, of course, to equalize the strain across each condenser of the series bank.

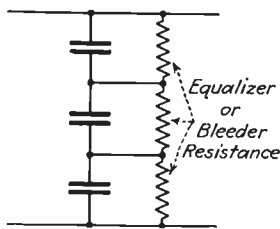


FIG. 366.—Bleeder resistance connection.

The resistance value of the equalizing resistance should be such that the current drain by the resistor is not excessive. A drain of 10 to 25 ma. is satisfactory in most instances.

Under any circumstances it is advisable to connect what is known as a *bleeder resistance* across the output of a rectifier unit, provided an equalizing resistor is not used. As before explained, this is to prevent excessive voltages from breaking down filter condensers, which might occur if the load were suddenly removed from the output. The load might be removed in a radio-telephone or telegraph transmitter if a tube in the load circuit burned out, or, in the latter, during keying. In a receiver or audio amplifier, the load might be removed if a tube burned out or was removed with the current on.

In power packs supplying power to receivers, the function of the bleeder resistance is performed by a section of the voltage divider, as will be explained when voltage dividers are discussed later. On the other hand, in rectifier units provided for transmitters, and in other circuits where no voltage divider is needed, the bleeder resistance is necessary not only to prevent condenser breakdown but to improve the regulation of the voltage output during the keying if a radio-telegraph transmitter is employed.

The value of the bleeder resistance in transmitting circuits depends on conditions in the circuit. A value of bleeder current equal to 25 or 30 per cent of the total drain on the rectifier unit is used in many installations with success. The value of bleeder resistance is calculated by Ohm's simple law, $R = \frac{E}{I}$, where E is the voltage drop across the resistor and I the bleeder current.

Bleeder resistances are sometimes variously known as *drain resistors*, *stabilizing resistors*, or simply as *load resistors*.

335. Voltage Dividers.—When the voltage has been rectified and filtered, it is ready to be applied to the vacuum-tube circuits of the transmitter or receiver. Except for simple one-tube circuits, it is nearly always necessary to apply different voltage values to various elements in the tube and to the various tubes or sets of tubes. In order to secure these different voltages, it is necessary to provide a separate power supply for each voltage or to divide the output of a single power supply into the various values desired. As it is usually uneconomical and unnecessary to provide a separate power supply for each voltage, one or two power supplies provide all the necessary voltages, for even a complicated radio circuit, when the output of the supply is divided by a voltage divider.

A voltage divider is very simple in its operation and functions entirely in accordance with Ohm's law. The development of an elementary voltage divider is shown in Fig. 367. In diagram (a), if the voltage is 100 volts and the current is 20 ma., the resistance R is found by the equation

$$R = \frac{E}{I} = \frac{100}{0.020} = 5,000 \text{ ohms}$$

It is seen then that, if the voltage and current are known, the value of the resistor R is easily found. Also, if any two of the values E , I , or R are known, it is simple to find the third by the application of Ohm's law.

Suppose now it is desired to tap off 40 volts from a resistor (voltage divider) across a 100-volt line. How much resistance must be used? The answer to this question depends on the load to be connected across the 40-volt line. For illustrative purposes let us assume that the load is to consist of a vacuum-tube circuit which draws 10 ma. Reduced to an elementary basis, the circuit appears as in Fig. 367 (b). In this circuit we have a resistance B in series with a parallel circuit consisting of resistance A and the tube load R . Now we know that the voltage drop across B must be the total line voltage, 100, minus the drop of 40 volts across the unit AR , or 60 volts. The next value we must have to progress with the calculation is the current value, I , which will flow through B . If we allow a bleeder or stabilizing current of 20 ma. through A , which

is the usual practice, and 10 ma. flow through R , a total of 30 ma. will flow through B . Applying Ohm's law we get

$$R_B = \frac{E}{I} = \frac{60}{0.030} = 2,000 \text{ ohms.}$$

The value of the parallel resistance AR is found, $R_{AR} = 40/0.030 = 1,333$ ohms. We know $R_R = 40/0.010 = 4,000$ ohms. Hence we must find out what value of resistance in parallel with 4,000 ohms will give 1,333

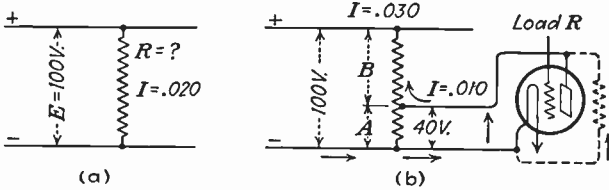


FIG. 367.—Fundamental circuit of voltage divider and tube-load connections.

ohms. For the two resistances A and R in parallel, the equation is

$$R_{AR} = \frac{1}{\frac{1}{R_A} + \frac{1}{R_R}} \text{ ohms.} \text{ Hence, if } R_R \text{ is known to be } 4,000 \text{ ohms,}$$

$$R_{AR} = \frac{1}{\frac{1}{R_A} + \frac{1}{4,000}} = 1,333 = \frac{1}{\frac{1}{1,333}} = \frac{1}{\frac{1}{3}} = \frac{1}{\frac{2}{4,000} + \frac{1}{4,000}}$$

$$R_A = \frac{4,000}{2} = 2,000 \text{ ohms.}$$

Proof: Since $R_{AR} = 1,333\omega$, and $R_B = 2,000\omega$, the total $R = 3,333$ ohms. Hence, the total E across AB will be

$$E = IR = 0.03 \times 3,333 = 99.99 \text{ or } 100 \text{ volts.}$$

The power expended in section A , which may be found by the conventional I^2R formula is $0.020^2 \times 2,000 = 0.8$ watt. The power expended in section B is $0.030^2 \times 2,000 = 1.8$ watts. There is, therefore, a total of 2.6 watts expended in the voltage divider. The power expended in the load R may be computed by the simple equation,

$$\text{Watts} = E \times I = 40 \times 0.010 = 0.4 \text{ watt.}$$

The total drain on the power supply is the sum of the power expended in the voltage divider plus the power expended in the load, 3.0 watts.

Unless there was a need for the 100 volts potential, this would hardly be an economical way to provide power for this load, as more power is wasted in the voltage divider than is used in the load. However, if a potential of 100 volts is needed for one purpose and a potential of 40 volts is needed for another purpose, it would probably be more economical to get the 40 volts by means of a voltage divider than to provide a separate

source of supply. It is situations of this kind that make a voltage divider desirable from an economical point of view.

It is well at this point to consider the type of resistors which would be suitable for the voltage divider explained. In section A the power expended is 0.8 watt. It has been found that resistors show better constancy of value and longer life if they are worked at but 50 per cent of their rated power-carrying capacity. Therefore, resistance A should have a rating of at least 1.6 watts. As a resistor of this capacity is not a stock item, it would be necessary to use a 2-watt resistor. Resistor B expends 1.8 watts, and it would be necessary to use a 4-watt resistor in this position, although a 10-watt size would be safer.

Any circuit in which a voltage divider is necessary can be reduced to a simple schematic diagram. Another example of this method is shown in Fig. 368. Assume that it is required to design a voltage divider to supply a number of tubes for a vacuum-tube receiver as indicated in the following list.

- Radio-frequency amplifier using two 235 screen-grid tubes.
- Detector using one 227 tube.
- Audio amplifier using one 227 tube.
- Power amplifier using two 245 tubes.

It will first be necessary to ascertain the plate and screen voltages and the total plate- and screen-current drain from the tube-characteristic table (see Appendix A, Table X). In this example the various voltages and total drain for each tube are as follows.

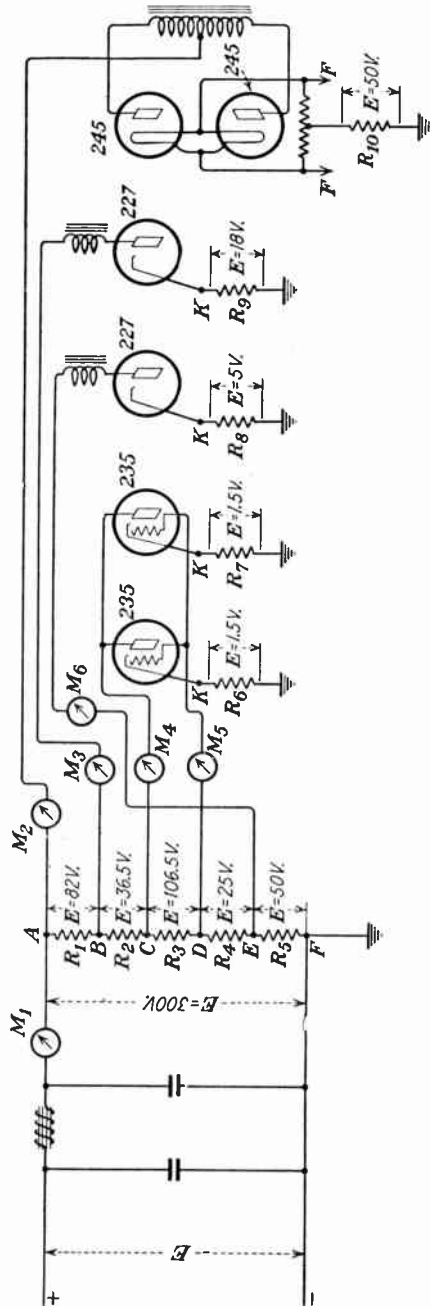


Fig. 368.—Typical voltage-divider circuit.

	Radio frequency 235	Detector 227	Audio frequency 227	Power 245
E_p	180 volts	45 volts	200 volts	250 volts
I_p	0.0058 amp.	0.0002 amp.	0.0051 amp.	0.034 amp.
E_c	1.5 volts	5 volts	18 volts	50 volts
E_{sg}	75 volts			
I_{sg}	0.0025 amp.			

The total drain for all tubes will be

Two 245	$0.034 \times 2 = 0.068$ amp.
One 227 (A.F.)	$0.0051 \times 1 = 0.0051$ amp.
One 227 (DETECTOR)	$0.0002 \times 1 = 0.0002$ amp.
Two 235 (R.F.)	$0.0058 \times 2 = 0.0116$ (plate) amp.
	$0.0025 \times 2 = 0.005$ (screen) amp.

Total drain = 0.0899 amp.

To this must be added a bleeder current of between 20 and 30 ma to accommodate increases in plate current due to signal-grid swings. This will insure a maximum degree of stability and prevent load variations in the power supply and its associated circuits.

In this example, we shall assume a bleeder current of 25 ma., or 0.025 amp.

Therefore the total drain to be delivered by the supply as indicated by the meter M_1 will be $0.0899 + 0.025 = 0.1149$ amp. From this value a current of 0.068 amp. will be delivered to the circuit M_2 . The remainder, or 0.0469, must then pass through the resistance R_1 . Before calculating R_1 , however, it must also be remembered that the total voltage across the divider between the points A and F must be 300 volts in order that 250 volts may be applied to the two power tubes. The value of 300 volts is required because there will be a voltage drop of 50 volts across the biasing resistance R_{10} . Now the drop across R_1 , necessary to decrease the voltage for the 227 audio-frequency tube to 200 volts, must be calculated on the basis of $200 + 18$ volts. The 18 volts being necessary to compensate for the biasing voltage lost across the resistance R_9 . Hence, to calculate the value of R_1 , we divide the drop necessary by the current flowing through it, $E_{AB} = 300 - 218 = 82$ volts. Since 0.068 passes through the circuit M_2 , the remaining current through R_1 must be $0.1149 - 0.068$ or 0.0469

$$R_1 = \frac{E}{I} = \frac{82}{0.0469} = 1,748^\omega$$

amp. Thus R_1 establishes an e.m.f. of 218 volts between the points B and F and an effective e.m.f. of 200 volts between the plates and filament of this tube.

Now, since the remaining current through R_2 to tap C is $0.0469 - 0.0051$ or 0.0418 amp., because 0.0051 is taken away by the circuit M_3 , and since R_2 must drop the voltage to 181.5, or a drop of 36.5 volts (1.5 volts being allowed for the drop in the biasing resistors R_6 and R_7),

$$R_2 = \frac{E}{I} = \frac{36.5}{0.0418} = 874^\omega.$$

The potential for the next tap, D , is then determined, since the screen grids in this circuit require a potential of 75 volts between D and F , or a drop of 106.5 volts.

The value $I = 0.0302$ is obtained by taking the difference between the current at point C and that taken by the circuit M_4 ; thus $0.0418 - 0.0116 = 0.0302$. Hence R_3 may be calculated thus:

$$R_3 = \frac{E}{I} = \frac{106.5}{0.0302} = 3,526\omega$$

Since the voltage for the detector tube from tap E must be 45 volts, the next resistance, R_4 , must drop the voltage to $45 + 5$ (R_5) or 50 volts, a drop of 25 volts.

The value $I = 0.0252$ is obtained by taking the difference between 0.0302 (at point D) and the current taken by the circuit M_5 ; thus $0.0302 - 0.005 = 0.0252$. Hence,

$$R_4 = \frac{E}{I} = \frac{25}{0.0252} = 992\omega$$

Now, since the last circuit M_6 will require a current drain of 0.0002 amp., the value of R_5 may be readily calculated. I is calculated: $0.0252 - 0.0002 = 0.025$

$$R_5 = \frac{E}{I} = \frac{50}{0.025} = 2,000\omega$$

The power dissipation for each resistance may then be calculated by the formulas I^2R or $E \times I$. The latter is preferable in this problem, since the voltages across each resistance have already been established.

IV. GENERAL RECTIFIER CONSIDERATIONS

336. Ripple frequency¹ is determined by the number of direct-current pulses delivered by the rectifier to the circuit per second. Assuming a 60-cycle current as standard, the ripple frequency for the various rectifier circuits before the filter is applied is as follows:

Circuit	Ripple Frequency, Cycles
Single phase:	
Half wave.....	60
Full wave.....	120
Full wave (bridge circuit).....	120
Voltage doubler.....	120
Three phase:	
Half wave.....	180
Half wave (double-Y).....	360
Full wave.....	360

The higher the ripple frequency, the easier it is to filter, that is, the smaller will be the values of L and C necessary in the filter.

The output wave form from a rectifier power supply depends on the effectiveness of the filter. Curve (d), Fig. 362, shows the wave form for

¹ The problem of ripple-filter design is beyond the scope of this text. Readers interested in this phase of the subject are referred to the references which follow:

DELLENBAUGH JR., F. S., and R. S. QUIMBY, *The Economical Design of Smoothing Filters*, QST, April, 1932.

JOHNSON, C. A., *Electric Filter Design*, Radio News, March, 1932.

an inadequately filtered supply, while curve (e) shows a well filtered wave form. To appreciate fully the effect of the ripple filter, curves (c) and (e) should be compared.

Multi-section filters are necessary if a single-section filter does not attenuate the ripple frequency sufficiently. A choke shunted by a condenser at each end is regarded as a single section of filter. Figure 364(a) is, therefore, a single-section filter preceded by a first choke. Diagram (c) is a single-section filter without the first choke, while diagram (b) is a two-section filter preceded by a first choke.

Filters Compared.—The output voltage from a choke-input filter is lower than if a condenser-input filter were used in its place. On the other hand, a rectifier equipped with a choke-input filter can be overloaded more without damaging the rectifier tubes than if a condenser-input filter was used. Of course, if mercury-vapor tubes are used, a choke-input filter *must* be used to protect the filament against excessive current surges.

337. Losses in Rectifier Apparatus.—The output voltage of the rectifier power supply is the voltage of the transformer secondary passed by the tube minus the loss in the rectifier tube and the filter. At first glance this might seem inconsequential, but it sometimes amounts to a considerable loss. As has been before pointed out, the loss in a high-vacuum type of rectifier depends on the amount of current being passed by the tube and may amount to a value of from only a few volts to several hundred volts. The loss in such a tube could be determined only by an inspection of the characteristic curve of the tube showing its voltage drop at various loads. The loss in a mercury-vapor tube regardless of the load is 15 volts under normal operating conditions. If two tubes are used in series, as in a voltage doubler or bridge circuit, the voltage loss is twice that of one tube.

The voltage loss in filter chokes may be figured by the formula $E = IR$. For example, a 15-henry choke with a resistance of 200 ohms passing 150 ma. would have a drop (loss) of 30 volts. If two such chokes are used in the circuit, the loss would be twice 30, or 60, volts. The power loss in each choke would be $I^2R = 0.0225 \times 200 = 4.5$ watts; in two such chokes the loss would be 9 watts.

338. Typical rectifier and filter systems are shown in the diagrams in the various chapters on receiving and transmitting apparatus. When the study of this chapter has been completed, these practical circuits should be studied with the view of thoroughly associating theory and practice.

Ripple filters for generators are described in the next chapter on Dynamo-electric Machinery. An inspection of the diagrams of transmitters using motor generators as a source of power will show the methods of connection employed and the values of condensers and chokes used.

339. Plate Fuses.—In the low- and intermediate-power transmitters, where circuit breakers or relays are not provided in the plate circuit, fuses may be inserted in the plate leads of the rectifier tubes (as well as in other tubes). The current-carrying capacity of these fuses must be such that they will afford the tube ample protection against overloads. Generally the normal load is from one-half to two-thirds the fuse rating. An inspection of any parts catalogue will give the reader an idea of the capacities of these fuses. They afford a very inexpensive insurance against tube damage from overloads.

340. Rectifier Troubles.—Troubles in the rectifier circuit show up as faulty circuit operation in either the transmitter or receiver for which the rectifier is the source of power supply. Perhaps the most common trouble is that of a faulty rectifier tube. If a high-vacuum type of tube is used, a faulty tube may be detected by a decrease in the output voltage of the rectifier. If a mercury-vapor tube is used, a faulty tube may be detected by the absence of the customary "healthy" blue haze appearing when the rectifier is in operation. The lack of a blue haze may be caused by air coming into the tube or by filament deactivation. Regardless of the cause, the tube will have to be replaced. It is important that the filament voltage of the rectifier be held constant at its proper value.

No output voltage from the rectifier may be due to faulty relays in the circuit the contacts of which need cleaning, or the armature of which needs bending to make it rest flat against the pole pieces. See that all relays are operating properly. The same holds true for door switches on transmitters, all of which must be closed for normal operation.

An incorrect load on the voltage divider caused by a ground or open circuit in equipment would cause trouble in the rectifier affecting its output and possibly burning out the voltage-divider section or sections affected.

Sometimes a type of sizzling interference called "hash" is encountered with mercury-vapor rectifiers. This trouble is caused by the sudden vaporization of the mercury vapor at the start of each alternation. The remedy is a grounded shield can placed over each rectifier tube and a radio-frequency choke in each plate lead at the rectifier-tube socket.

A common cause of trouble in a rectifier system is the blowing out or breaking down of a condenser. This is usually due to excessive current surges which strain the condenser above its working voltage point. To guard against condenser breakdown, condensers with a working voltage well above the maximum surge voltage must be used.

A defective condenser may be detected by a continuity test using a voltmeter and battery in series across the condenser terminals. If the condenser is in good condition, the voltmeter will show a temporary deflection when the voltage is applied across the condenser showing that the condenser is absorbing a charge. When the condenser is charged the

deflection drops to zero. If the condenser is short-circuited by connecting together its terminals after this test is made, a flash will occur as the condenser discharges just before the connection is made. This is further proof that the condenser is in good condition. If the condenser is short-circuited internally, owing to breakdown of the dielectric by excessive potential, a continuous deflection will be noted on the voltmeter.

A defective choke coil could be located by a similar continuity test, using a voltmeter and battery. An open choke coil would, of course, be indicated by no deflection on the voltmeter. A choke with part of its winding internally short-circuited would be indicated by a higher reading on the voltmeter than is secured by comparison with a perfect choke of the same type and size. Troubles in rectifier circuits can usually be located by continuity tests using a voltmeter or buzzer in series with a battery across the parts to be tested.

CHAPTER XIV

DYNAMO-ELECTRIC MACHINERY AND METERS

Radio transmitters are so dependent upon the power supplied by dynamo-electric machines that a study of motors and generators becomes a requisite in any consideration of radio-communication systems. In this chapter, therefore, applications of the dynamo-electric principle to modern power-supply apparatus will be examined.

I. GENERATORS

A generator is a machine for converting mechanical energy into electrical energy. It must be driven by a prime mover which may take the form of a hand- or wind-driven machine, a steam, gas, or oil engine, or an electric motor, the combination being called a *motor-generator* which machine is commonly employed in radio practice.

341. Types of Alternators. *a. Introduction.*—The principles of operation of the simple alternator were fully treated in Sec. 41. The practical alternating-current generator or *alternator* is made in three distinct types, (1) the revolving-armature type, (2) the revolving-field type, and (3) the inductor type. The revolving-armature type of machine is usually a low-power machine owing to the difficulty of taking off heavy current from the revolving armature. Some of the old-type spark transmitters, such as the Marconi type P-4 or P-8, later converted into a tube transmitter and called the RCA type ET-3628, employ a G.E. 2-kw. revolving-armature-type generator. This type of machine has been almost wholly superseded by the inductor-type machine for low-power radio requirements. High-power alternators for power purposes are generally of the revolving-field type.

b. The revolving-armature type has the field poles fastened directly to the frame which, in turn, is bolted to the body casing or bedplate of the generator unit. The armature revolving in this field generates in itself an e.m.f. of alternating characteristics in a manner identical with the action of the simple loop in Fig. 49 in the chapter on Alternating-current Electricity. Current is taken from the machine by two slip rings and brushes as shown in the schematic construction diagram, Fig. 369.

c. The revolving-field type is generally used in large alternators in which high voltages are to be generated. This is shown in Fig. 370. Here the rotating field coils are excited by a constant e.m.f. and an alternating e.m.f. is induced in the stationary-pole windings as soon as

the field is rotated. Note the difference: in the stationary-field-type alternator the alternating e.m.f. is generated in the rotating armature winding and taken out to the external circuit by collector rings, whereas in the revolving-field type the alternating e.m.f. is taken out from the

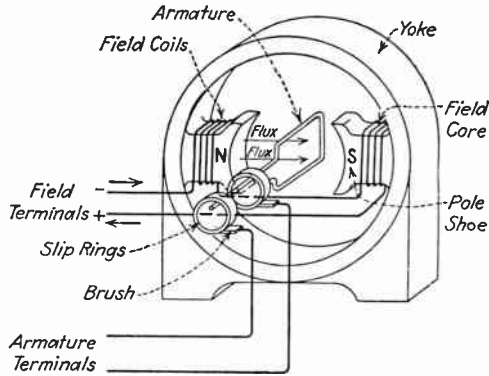


FIG. 369.—Mechanical arrangement of simple alternator.

stationary-pole windings, now the armature, when the rotating field is excited by a direct current through collector rings. The student will note that in either type of alternator the theory of induction holds true, that is, an e.m.f. may be induced into a conductor if the conductor is moved in a magnetic field, or an e.m.f. may be induced into a conductor if the conductor is placed in a moving magnetic field.

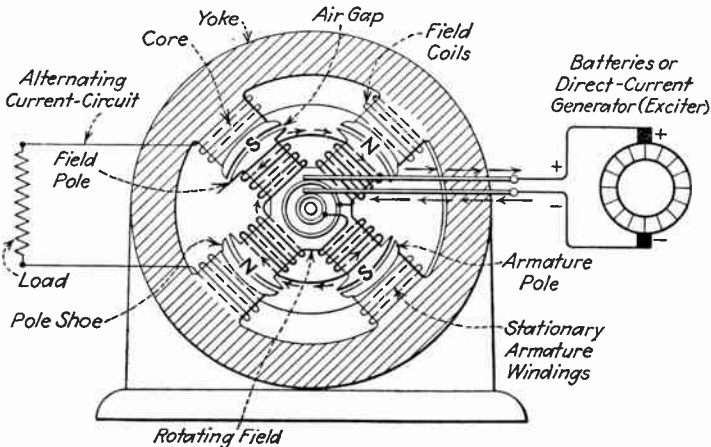


FIG. 370.—Revolving-field-type alternator.

In the first type of alternator the alternating e.m.f. is generated in the armature if it is revolving in a properly excited stationary field, and the alternating current can then be taken from the armature by collector rings and, similarly, an alternating current can be taken directly

from the stationary winding of the second type of alternator when the field is rotating around it, provided the rotating field is properly excited as in Fig. 370. In other words, the stator coils take the place of the rotating armature in this type and might be referred to as a *stator armature*.

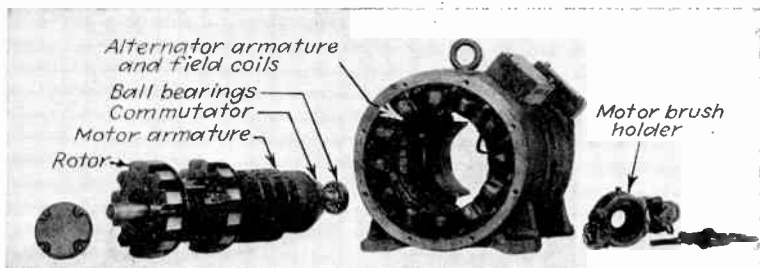


FIG. 371.—Parts of inductor-type alternator.

d. The Inductor Type.—In the inductor alternator the armature and field windings are fixed, but a revolving rotor of steel with toothed projections rotates in the magnetic path of both windings. This rotor

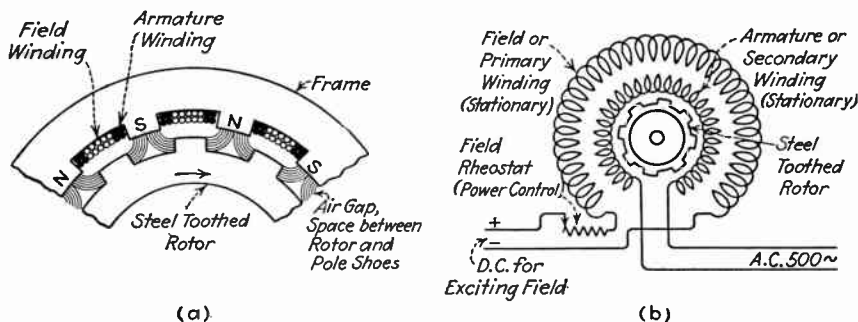


FIG. 372.—Inductor-type alternator: (a) how rotor cuts flux; (b) schematic diagram.

is a solid mass of steel and carries no winding (Fig. 371). The theoretical action can be clearly understood from Fig. 372(a). The rotor when revolving in the excited magnetic field periodically varies the reluctance, and thus produces a constant increasing and decreasing effect upon the magnetic flux. Therefore, when the flux is increasing, the induced e.m.f. will have a certain polarity. Then, upon decrease, the induced e.m.f. reverses and an alternating e.m.f. is generated. This type of design is used in practically all of the 500-cycle generators used with spark transmitters. A schematic diagram of the inductor-type alternator is shown in Fig. 372(b).

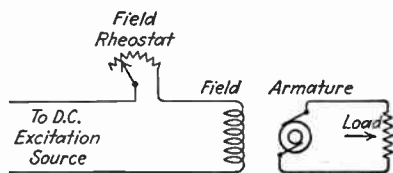


FIG. 373.—Field rheostat (voltage control) connections on alternator.

342. Voltage Control on Alternators.—

As most alternators are separately excited, it is a simple matter to regulate or control the output voltage by varying the excitation current. This is done by inserting a variable resistance in series with the field of the alternator, as shown in Fig. 373. Such a control device is known as a *field rheostat*. Field rheostats for use with alternators usually have a resistance equal to about twice the resistance of the alternator field.

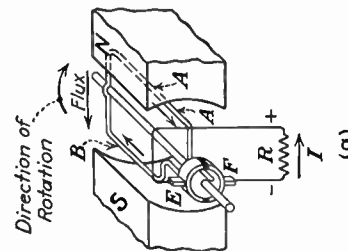
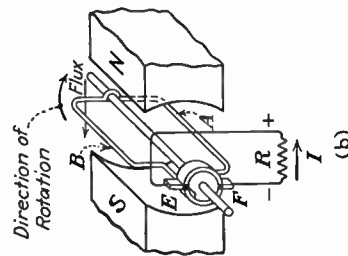
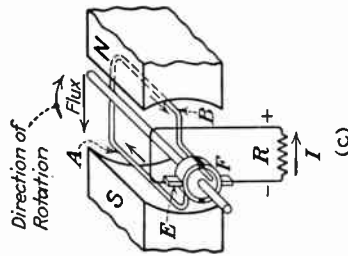
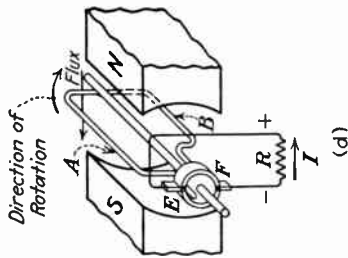
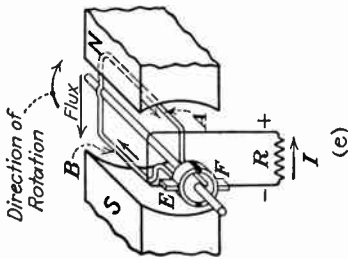


Fig. 374.—Commutator action through one complete revolution of the armature.

343. Direct-current Generators.—

In many respects the direct-current generator is similar to the alternator, the principle of current generation in both machines being the same, that is, an e.m.f. is induced in the armature when the armature conductors are cut by the flux produced by the field of the machine. Also the e.m.f. induced in the armature of the direct-current machine is dependent upon the flux density in the field and the speed of cutting.

There are, however, two points of notable difference between the alternating-current and direct-current machine and these are: (1) the alternator may have a stationary or rotating field, or it may be an inductor-type machine, whereas the direct-current machine is always a rotating-armature machine; (2) the alternator is always fitted with slip rings (except the inductor-type machine), whereas the direct-current machine is always fitted with a commutator, the function of which will be explained later.

The parts of the direct-current machine are the *field*, the *armature*, and the *commutator*. The field is always stationary, the armature is always rotating, and the commutator is always

necessary. The magnetic circuit of the direct-current machine is the same as that of the alternating-current machine, and the names of the magnetic circuit parts are similar.

Whereas alternating-current machines are nearly always separately excited, the direct-current machine is nearly always self-excited. For the same reasons that alternating-current machines are usually multi-polar so, too, are direct-current machines multi-polar. The current and voltage characteristics of the direct-current machine are very simple and follow Ohm's law as explained in Sec. 9.

It is evident from this summary of the similarities between the two machines that much of the theory of alternating-current machines also applies to direct-current machines. There are, however, some points of great difference which require special study. These points will now be explained.

344. Commutation.—It was seen in our study of the generation of an e.m.f. in a rotating loop that the current flow in the loop reversed itself once during each revolution of the loop. This caused the current to flow in a positive direction, let us say, when the loop was rotating in the arc from 0 to 180 degrees, and in the negative direction when the loop completed the revolution from 180 to 360 degrees. There was, therefore, a forward and backward flow for each revolution of the armature in Fig. 369 and this constituted the simple alternator.



FIG. 375.—Small size commutator.

Now, in order to get a direct current at the output terminals of the machine, it is necessary to rectify this alternating current generated in the armature. This is accomplished in a generator by a commutator which, in the simple machine about to be explained, consists of two semicircular current-collecting plates called *commutator bars* or segments fitted to one end of the armature as shown in Fig. 374. A commutator for a practical machine would have many bars, such as shown in Fig. 375. Large machines may have several hundred bars, but there are generally between 20 and 30 bars on small machines built for 110-volt circuits.

The function of the generator commutator is to reverse the connections of the armature conductors to the external circuit coincidentally with a change of induced e.m.fs. in these conductors, as a study of commutation will show. If the loop in Fig. 374(a) is rotating in a clockwise direction, as indicated by the arrow, an e.m.f. will be induced in the loop and a current will flow in a direction along conductor A which would be toward the reader and along conductor B in a direction away from the reader. This is in accordance with the left-hand rule for revolving-armature machines explained in Sec. 36 on Production of E.M.F. by Field Cutting Conductor. From the conductor A the current flows through the external circuit R in the direction of the arrow. It is conducted to this circuit through the commutator bars attached

to the conductors *A* and *B*, through the brushes making contact with these respective bars, and thence to the external-circuit line. Notice that the top brush is the positive terminal and the bottom brush the negative terminal of the generator.

Now, when the loop has made a quarter of a revolution [Fig. 374(b)], the conductors are running parallel with the magnetic field, are not cutting it, and, therefore, no e.m.f. is induced in the loop. At this moment the brushes are making contact only with the insulated portion of the commutator while they are about to pass from one segment to the next, or, in this particular machine, to the opposite segment. When this is done [Fig. 374(c)], the connections to the external circuit from the conductors have been reversed, that is, conductor *A* is now connected to the top brush and conductor *B* to the bottom brush. In the first quarter of a revolution completed by the loop, the e.m.f. induced in the loop might be graphically shown to be from point *a* to point *b* on the voltage curve of Fig. 376.

As the rotation continues through the second quarter, conductor *A* proceeds to pass up in front of the south pole with the result that the current in this conductor is caused to flow in the opposite direction to the way it flowed when this conductor was passing the north pole. Also notice that conductor *B* is passing downward in front of the north pole, and that the direction of current flow is opposite to what it was when this conductor was passing up in front of the south pole [see Fig. 374(c)]. By carefully following the explanation just given, it will be seen that the current flow in the

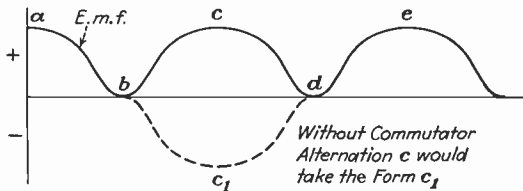


FIG. 376.—Effect of commutator on alternating current flowing in generator armature.

loop has reversed itself, and, as far as the loop alone is concerned, an alternating current has been generated. However, the commutator has also reversed the connections of the loop to the outside circuit, that is, conductor *A* is now connected to the top brush and conductor *B* to the bottom brush. These connections and direction of current flow continue as long as conductor *A* is passing up in front of the south pole and conductor *B* down in front of the north pole, or to the end of the third quarter.

When the third quarter has been completed, the brushes again pass over the insulated section of the commutator from one bar to the next, reversing again the connections from the conductors to the external circuit. During the last quarter, conductor *A* is again connected to the bottom brush and conductor *B* to the top brush. Conditions are now the same as they were in the first quarter, and the revolution has been completed.

On the voltage curve (Fig. 376), point *b* to point *c* is the e.m.f. generated during the second quarter; point *c* to point *d*, during the third quarter; point *d* to point *e*, during the fourth quarter. Therefore, with a complete revolution of the loop, two pulses of direct current are sent into the external circuit. As the revolution of the armature continues, the pulses continue to flow in the external circuit and a direct current is being generated. Of course, in a practical machine there are many armature turns or loops, and many pulses are generated in a revolution giving the effect of a steady current of unvarying amplitude in the external circuit.

It is seen from this elementary explanation that the current in either conductor of the loop of the simple machine, shown in Fig. 374, is in a direction toward the reader when either conductor is passing down in front of the north pole; and, away

from the reader, when either conductor is passing up in front of the south pole. Also notice that the top brush is always connected to the conductor passing the south pole, and that the bottom brush is always connected to the conductor passing the north pole. As the direction of the current flowing in the conductor to which the top brush is connected at any given moment (except when the brushes are passing from one bar to the next over the insulated mica-segment portion of the commutator) is always the same, it follows that a direct current must flow in the external circuit.

The principle of commutation just explained applies to all direct-current machines regardless of size or number of bars in the commutator.

345. Commutator Sparking.—When there is no current in the armature circuit, and consequently no armature reaction or cross-magnetization, the field of the machine is in a normal position as shown in Fig. 377(a). As soon as a current starts to flow in the armature, the field of the machine is shifted owing to *armature reaction* as shown in Fig. 377(b). The neutral axis has been moved from line AB to CD which shows that, for a generator, armature reaction tends to move the neutral axis in the direction of the rotation of the machine. In a motor this tendency is reversed, and the neutral axis is moved against the direction of rotation. For sparkless commutation the brushes must be set in the neutral axis. The reason for this is as follows:

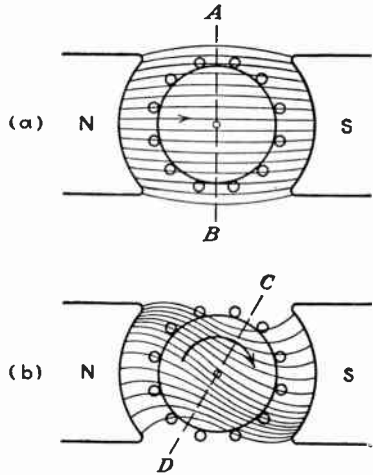


FIG. 377.—Effect of armature reaction.

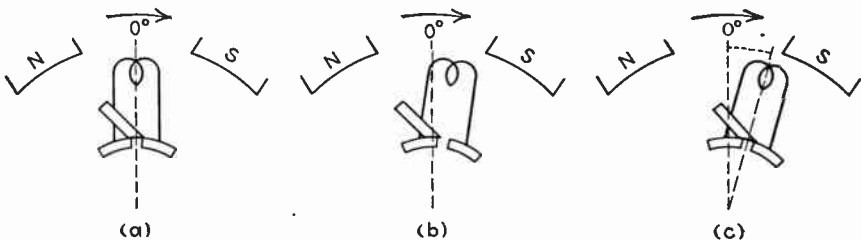


FIG. 378.—Graphical analysis of brush sparking.

In Fig. 378(a) is shown an armature element halfway between two adjacent poles. This is the position of neutral field when there is no field shift due to armature reaction, an effect which need not be considered as affecting this explanation. As the coil is passing through the position of neutral field, no e.m.f. is being generated or induced in the coil. Furthermore, the brush short-circuits (by-passes) the coil, and no current flows from it. However, the self-inductance of the coil tends to maintain a current flow in the coil even though it is short-circuited. This current, by the way, is a north-pole-induced current and as such flows in a certain direction, which for the purpose of this explanation we shall call clockwise.

As the coil comes under the influence of the south pole (b), which induces a counterclockwise current in the coil, there is a very sudden tendency for the clockwise current which was due to the north pole to cease flowing. At this same instant the brush is breaking contact with the bars and ceasing to short-circuit them. It might be said at this point that the brush did not short-circuit the coil long enough for the clockwise current entirely to cease flowing, and, therefore, at the moment the brush is breaking contact with the bars, there is a relatively high current set up in the coil (by the sudden reduction of the clockwise current by the counterclockwise current), which causes an arc to form between the brush and the bar it is leaving. This arc constitutes what is commonly referred to as *sparkling* and, besides being a direct power loss, it soon ruins the commutator.

In small machines this sparking may not be severe, but in heavy-duty machines a method of reducing sparking at the brushes must be adopted if the machine is to deliver its rated output.

Sparking could be reduced by moving the brush forward (in a generator) to the commutating axis denoted by the dotted line to the right of 0 degrees as indicated in (c), but as this would result in the coil being short-circuited, when it is cutting considerable flux, a loss of useful power would result, and the machine would not deliver its rated output.

When the brushes are moved forward in the direction of rotation, the brushes are said to have a *lead*. When they are moved counter to the direction of rotation, they are said to have a *lag*. In a generator the brushes must lead slightly for sparkless commutation, while in a motor they must lag unless the machine is fitted with interpoles in which case the brushes are fixed at the neutral point at all times.

346. Commutating Poles.—The problem of brush sparking has been solved by the use of *commutating poles* which are also known as *interpoles*. As shown in the preceding explanation of brush sparking, this sparking is caused by the self-induction of the armature coil in contact with the brush. In order to counteract this self-induction it is necessary quickly to reduce the current in the armature coil as it passes between poles and to cause a reversal of the current while it is yet being short-circuited by the brush so that the reversal will not come so suddenly, just at the moment when the brush is leaving the bar, and thus cause abnormal sparking. This can be accomplished by placing a small field pole between each main field pole of the machine. These commutating poles or interpoles consist of a relatively few turns of heavy wire forming a coil which is connected in series with the armature, as shown in Fig. 379(a). It is to be noted that the polarity of the interpole of a generator is that of the next field pole, in the direction of rotation; while in a motor the polarity of the interpole is that of the field pole just passed, in the direction of rotation as shown in (b). The reason for this difference will be clear when the theory of motor rotation is understood which will be explained in Sec. 356 on motor principles.

The advantages of the commutating poles are as follows: Without interpoles the brushes must be shifted to the neutral point, which varies with the load; with interpoles the brushes are always at the neutral point owing to the action of the interpoles in counteracting the self-induction of the armature; without interpoles the momentary overloads are limited to about 50 per cent of the rated capacity of the machine, but

with interpoles these overloads may range as high as 200, 300, or even 400 per cent without damage to the machine owing to the absence of sparking; motors without interpoles are limited in their speed range to 40 or 50 per cent above normal; motors with interpoles may have their speed increased four or five times normal without harm.

In addition to the points already mentioned, interpoles neutralize armature reaction, and the losses consequent thereto, and raise the overall efficiency of the machine to a considerable and important degree.

347. Compensating-field Windings.—As the size of the direct-current motor or generator is increased so is the armature reaction, and the consequent field distortion. In order to reduce this field distortion, a few turns of heavy conductor are wound around the face of each field pole to provide a m.m.f. equal but opposite to that produced by the armature.

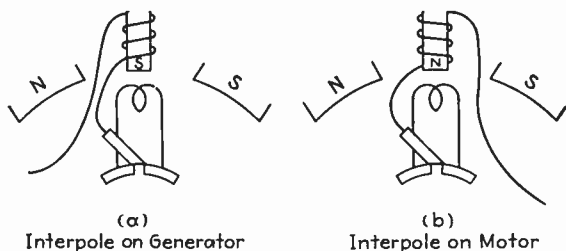


FIG. 379.—Interpole connections.

These *compensating turns* on each pole form a *compensating coil*. The compensating coils on all the field poles are connected in series to form the *compensating-field winding*. This compensating-field winding is then connected in series with the armature, as in this position the compensating field is directly affected by the load on the machine. When the load is light, the armature reaction is light. The armature reaction being light, it requires only a small compensating field to oppose its distorting effect on the main field. This is automatically taken care of by the lighter field produced by the light current flowing through the compensating field which is in series with the armature drawing a small current on a light load. Likewise, when the load is increased, so is the compensating field.

Commutating poles as well as compensating fields may be used on the same machine, but when compensating fields are used, the ampere turns on the commutating poles may be reduced as much as 70 or 80 per cent compared with the turns necessary without the compensating fields.

It should be understood, however, that compensating fields do not aid the reversal of the current in the short-circuited armature coil thereby aiding commutation; but rather the aid to commutation lent by the compensating fields comes through their opposition to armature reaction and the consequent reduction in field distortion thereby affected.

348. Double-commutator Generators.—The weakest point in a high-voltage direct-current generator is the armature insulation. When an output of over 1,000 volts is generated, it becomes extremely difficult to insulate the armature windings and commutator and to protect against high-voltage breakdown. In spite of this difficulty, generators are commercially available to supply as high as 12,000 volts. On these high-voltage generators it is the practice to divide the output-voltage strain between two commutators, allowing each commutator to take half the strain. To get full voltage, the outputs from the two commutators are connected in series. Spare armatures are usually available wherever these high-voltage machines are used so that, if an armature does break down, it may be quickly replaced.

FIG. 380.—Skeleton view of drum armature.

349. The drum armature is so called because the windings are wound over the outside of the armature core as over the outside of a drum, all of the wire laying on the surface of the core (see Fig. 380). The idea of the drum winding is that the two conductors of each loop (remember, a practical armature is similar theoretically to the simple armature shown in Fig. 374, only it has more than one loop) shall lie under adjacent unlike poles. Study this idea and the construction of armatures will become simple.

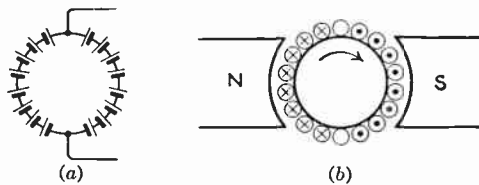


FIG. 381.—Armature equivalents.

Now let us assume, for example, that an armature has 100 loops, they are connected in series. Electrically this armature is equivalent to a number of cells in series as shown in Fig. 381(a). It was shown in our previous study of the revolving loop that conductors lying under unlike poles have unlike polarities. Therefore, the conductors on an armature with many loops would have polarities as indicated on the end views of the conductors, Fig. 381(b). In the conductors marked with a cross, the current is flowing toward the reader; in those marked with a dot, away from the reader. The external appearance of most direct-current armatures is the same, but there are two distinct types of drum windings. There are the *lap* winding and the *wave* winding. These are shown

diagrammatically in Fig. 382(a) and (b). A close study and tracing out of the electrical path in these armatures will make the wave and lap types of windings clear.

In low-power machines each winding element consists of more than one loop, in fact it consists of several turns which taken together are called an armature coil. All of the armature coils in an armature are

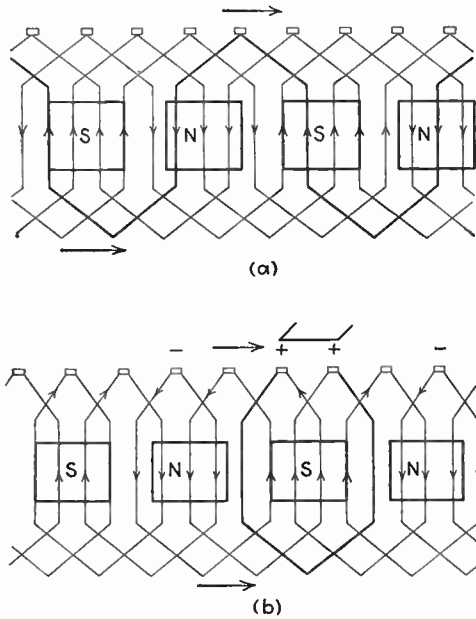


FIG. 382.—Armature windings: (a) wave winding; (b) lap winding.

alike and are interchangeable so that a damaged coil may be removed and replaced. Each turn that goes to make up an armature coil is carefully insulated with mica and cotton, or linen tape. They are then fastened together into a coil or *form* by wrapping the coil in mica and a few layers of insulating tape, the whole coil then being dipped into a moisture-proof varnish or treated by a vacuum impregnation.

350. Armature Cores.—The core for the modern drum armature is made up of annealed steel punchings shaped as shown in Fig. 383. These punchings are placed one on top of another to form a cylinder of laminated sheet steel. A key slot is provided in the center hole to prevent rotation after the punchings are assembled on the axle. The indentations around the perimeter of the punching form the slots in which the armature conductors are placed.

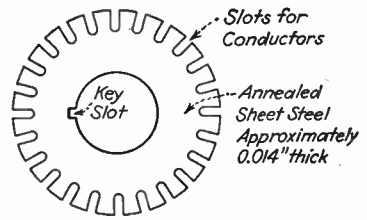


FIG. 383.—Steel armature-core punching.

From the design standpoint, it is necessary to build generators for radio service with as large a number of armature slots as possible, to decrease the amount of slot ripple and increase its frequency, and also to furnish as large a number of commutator bars as possible, in order to keep down the voltage between the bars and thus reduce the possibility of sparking as well as increasing the frequency of the commutator ripple. It is easier to filter a ripple when its frequency is high; hence the desire to increase ripple frequency.

351. Field Excitation. *a. Requirements.*—In our study of the alternator we saw that, in order to produce a field flux of sufficient

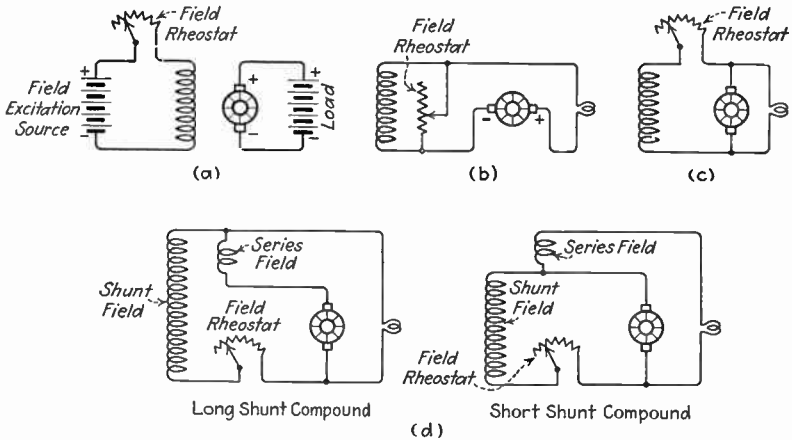


FIG. 384.—Methods of field excitation: (a) separately excited; (b) series excitation; (c) shunt excitation; (d) methods of compound excitation.

density to make practical generation of e.m.f. possible, it is necessary that the field magnets be excited. It was also shown that with alternators it is necessary to separately excite the machine. It is, conversely, the general practice to make direct-current generators self-excited. An exception to this rule is a direct-current generator designed for storage-battery charging purposes, or for electroplating work, in both of which instances the machine is generally separately excited.

An elementary diagram of a *separately excited* machine is shown in Fig. 384(a). The load on this machine is a storage battery which is being charged. The field-excitation current may be a storage battery as shown, or any other source of direct current. Notice that the armature is drawn showing only the commutator. This is the conventional method of showing a direct-current machine.

b. Series excitation is shown in Fig. 384(b). When the armature of this machine starts to turn over, it cuts the lines of force set up by the residual magnetism of the field poles which induces an e.m.f. in the armature. If now the armature continues to rotate and the circuit is closed, a current will flow. This current flows out of the armature, through the field and load, and back to the armature. The current in the armature circuit, which is compelled to flow through the field circuit, excites the field,

and the machine is said to be self-excited. Because the entire current generated by the series machine must pass through the field winding, this winding must be of a heavy wire of sufficient current-carrying capacity to handle the load without undue heating.

c. Shunt excitation is shown in Fig. 384(c). As soon as the armature starts to rotate, the residual field induces an e.m.f. in the armature which tends to increase the field flux. As the rotation continues, the e.m.f. in the armature increases and, as the field is connected across or in *shunt* to the armature, a current will flow. If a load is connected to the armature as shown in the figure, the armature current will divide itself, part flowing in the field, and part to the load, circuits. The field winding of a shunt machine may consist of several thousand turns of insulated wire, the exact number of turns depending on the size of the machine. This field winding, therefore, has a high resistance and allows only a small fraction of the armature current (3 to 5 per cent) to flow through the field; the rest flows through the load circuit. It is evident that the shunt machine is self-excited by virtue of its field being connected across the armature. Because of the relatively small current flowing through the field of a shunt machine, the shunt windings are of a small-size wire compared with series-field windings.

d. Compound excitation is a combination of series and shunt excitation as will be seen by referring to the diagrams in Fig. 384(d). It will be noted that there are two methods of connecting the shunt field, that is, it may or may not be shunted around the series field. If the shunt field is connected around the series field it is called a *long shunt*; if not, it is called a *short shunt*. Usually, when a compound-wound machine is referred to, a long-shunt machine is meant unless specifically stated otherwise.

In a compound machine the series field consists of relatively few turns of heavy wire, while the shunt field consists of many turns of fine wire. All field windings are of insulated wire, of course.

Compound generators are always *cumulatively wound*, that is, the *series and shunt fields aid each other* in making the total flux of greater density than that of either field alone.

352. Brushes.—The brushes of a direct-current machine make contact with the commutator. Brushes are made in two types, namely, *copper-leaf* brushes and *carbon* brushes. The former have a current-carrying capacity of 150 amp. per square inch and are, therefore, used on heavy-duty machines. Carbon brushes can carry 35 amp. per square inch and are satisfactory for small machines.

A multi-polar machine may have only one pair of brushes, or more than one pair might be used. The number of brushes used never exceeds the number of poles on the machine. The reason for more than one pair of brushes on a machine is to divide the load and reduce commutator sparking, heating, and other excessive wear which tend to ruin the commutator. On a two-wire system, if more than one pair of brushes is used on the generator, all of the positive brushes are connected to one terminal of the machine, and all of the negative brushes are connected to the other terminal.

353. Voltage Control on Direct-current Generators.—The terminal voltage of separately excited, shunt, and compound generators is controlled by connecting a variable resistance called a *field rheostat* in series

with the field coils, as shown in Fig. 384. The terminal voltage of a series machine may be controlled by shunting a variable resistance of high current-carrying capacity around the series field thereby controlling the amount of current flowing through the field and consequently affecting its excitation.

354. Uses of Various Types of Generators.—Whether a generator should be separately, series, shunt, or compound excited depends entirely on the service demanded. In addition to the characteristics pointed out before, the separately excited generator will show a variation in terminal voltage almost in direct proportion to the speed variation; that is, the voltage rises in a straight-line characteristic as the speed is increased and vice versa. The effect of varying the speed of shunt and compound generators is greater than in the separately excited machine. In other words, this characteristic is expressed as being steeper. This means that the voltage rises more quickly in shunt and compound machines as speed is increased and also falls more quickly when speed is reduced than does the voltage of the separately excited machine.

Where the load does not change much, a simple shunt machine is satisfactory. The shunt generator, therefore, is used extensively for battery-charging purposes and as an exciter for alternators. On the other hand, where there are sudden and large changes in load to take care of, the compound-wound machine is more suitable.

355. Regulation.—When a load is applied to a generator there is a drop in the output voltage from a *no-load* (*nl*) value to a *full-load* (*fl*) value. This change in voltage under these conditions is called the *inherent* regulation of the machine. If expressed as a percentage of the rated *no-load* voltage by applying the following equation,

$$\text{Regulation percentage} = \frac{E_{nl} - E_{fl}}{E_{fl}} \times 100. \quad (178)$$

Regulation as expressed above is dependent upon the power factor of the circuit. A lagging current causes a greater reduction in voltage when the load is applied than the same current would cause were it in phase with the voltage. This, of course, affects regulation adversely. On the other hand, a leading current, such as is caused by a capacitive load, improves the regulation, causing a higher terminal voltage with full load than with no load.

The term *regulation* meaning inherent regulation in per cent is ordinarily applied to generators, motors, transformer, rectifiers, filters, and rectifier-filter combinations. Voltage regulation is usually meant in all references except to that of the motor when speed regulation is the factor of interest. The equation here given lends itself to application to any of these devices or factors. For example, when figuring the regulation of a motor generator, either speed, frequency, or voltage regulation

could be computed by letting the subscript *nl* stand for no-load conditions, and the subscript *fl* represent full-load conditions when adjoined to the appropriate letter to represent speed (*S*), voltage (*E* or *V*), or frequency (*f*). When figuring transformer regulation, only voltage is involved.

The speed of a generator affects its regulation, owing to the induced e.m.f. being dependent upon the rate of cutting of the lines of force by the armature conductors, assuming that the field excitation remains fixed. In other words, as the speed is varied, the terminal voltage is varied proportionately. Best regulation is obtained when the machine is run at its normal speed.

Full load in the above equation as represented by the subscript *fl* means the full load to which the regulation applies. For example, the regulation of a motor generator or rectifier may be 25 per cent at half load, or it may be 35 per cent at maximum load. However, if the apparatus is used on a transmitter which draws only one-half the power for which the apparatus is designed, full load is only 50 per cent of the maximum rated load that the apparatus is capable of handling. Therefore, for our equation the full-load factor is the load which is on the power supply when the key is pressed. If the same power supply, be it a generator or rectifier, is used on another transmitter which draws the maximum rated load of the device, then a new value of regulation percentage would be found. Hence, when a regulation percentage is stated or referred to, it is assumed to apply to the maximum load on the apparatus, unless it is specifically stated otherwise.

II. MOTORS

An electric motor is a device for converting electrical energy into mechanical energy. In this respect its function is just the reverse of that of the generator. A motor will produce mechanical energy only when an external source of electrical energy is forced through its windings by an impressed voltage.

356. The construction principles of the direct-current motor are the same as those of the direct-current generator, and the motor may, in nearly every instance, be used as a generator. The basic principles on which the motor operates are explained as follows:

When a wire carrying no current lies at rest in a parallel magnetic field, the magnetic lines of force pass from the north to the south pole undisturbed, as shown in Fig. 377. As soon as a current flows in the wire, this condition is changed, for a field immediately surrounds the wire. The lines of force constituting this field rotate counterclockwise looking along the wire in the direction of current flow. As the conductor is in the midst of the field produced by the poles, it is evident that on one side of the wire the field around the conductor and the field between the poles of the machine will counteract one another thinning out the field, while on the opposite

side of the conductor these two fields will join, being both in the same direction, strengthening the field, and forcing the conductor sideways.

When there are many armature conductors, as there are in all practical motors, the effect of the field produced by the armature conductors on the field produced by the field poles is as shown in Fig. 386, where the entire field is distorted, the lines being much denser on one side of the conductors than on the other. This uneven density of the field flux causes the armature to rotate.

The armature of a direct-current motor is exactly like that of a direct-current generator and may be built in accordance with any of the conventional methods of building direct-current machines as explained in Sec. 349.

Based on the fact that *actual current flow in a wire is from negative toward positive*, a rule called the *right-hand rule* has been formulated to

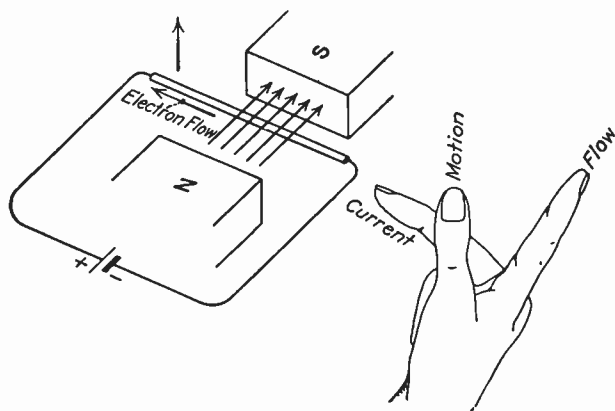


Fig. 385.—Right-hand rule for motors.

give the relationship between current flow in the armature, direction of field flux, and direction of rotation in a revolving armature direct-current motor. This rule is stated as follows (see Fig. 385): *Point the forefinger in the direction of the flux, the middle finger in the direction of current flow; then the thumb will point in the direction of rotation.*

357. The function of the motor commutator on a direct-current motor is to maintain the proper direction of current flow in the armature coils in relation to the magnetic field, so that the rotative effort of all coils will be in the same direction. All direct-current motors are equipped with commutators. Commutators are also used on certain types of alternating-current motors for starting purposes.

When an external source of e.m.f. is applied to the terminals of a direct-current motor, a current is forced through the armature and field windings. The current which passes through the field windings sets up a magnetic field between the pole faces. The armature conductors lying in this magnetic field and carrying current are forced to move out, and

other conductors are carried in between the pole faces. The current flowing in the conductors just entering the magnetic field is kept in the same direction as that in the conductors leaving the field, by means of the commutator, which reverses the current in the armature coils at the instant the coil is at the neutral point between the poles. Figure 386 shows the attraction and repulsion between the magnetic fields and the conductors of the armature in a motor.

358. Counter Electromotive Force : Torque.—There are two e.m.f.s. present in a motor armature: the greater is the impressed e.m.f. which is sent into the armature from an outside source; the other is an e.m.f. induced in the armature conductors when it rotates in the motor field.

This last mentioned e.m.f. is called a *counter e.m.f.* because it is of opposite polarity to the impressed e.m.f. As the direct-current resistance of the armature is fixed in value by the amount and size of the armature-winding wire, and as the impressed voltage is normally fixed, the current flowing in the armature is governed by the counter e.m.f. produced by the armature rotation.

If the speed of the motor is increased, the counter e.m.f. increases, the drop across the armature decreases, and the current decreases. It is for this reason that a motor draws more current on starting than after it is up to speed; there is no counter e.m.f. in an armature which has not yet commenced to rotate.

The counter e.m.f. is always just a little less than the impressed e.m.f. in a motor.

Torque is the measure of the tendency in a body to rotate, and it may be expressed in foot-pounds of force at a given radius. It is not necessary that there be motion in order that there be torque. As an example, suppose the emergency brake on an automobile is tightly set so that the rear wheels cannot possibly move. If power is now applied (not enough to cause the brakes to slip), a force is exerted on the wheels although there is no motion. Torque is the twisting power exerted on an object. It is easily seen that an enormous torque may be applied and not a single foot-pound of work done. Torque in an electric motor varies as the product of work done. Torque in an electric motor varies as the product of the field density (flux), times the armature current.

359. Types of Direct-current Motors.—Direct-current motors have their fields and armatures connected in various ways similar to the methods employed for direct-current generators. This can be seen by comparing the generator connections shown in Fig. 384 with the motor

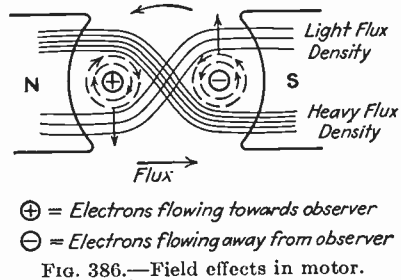
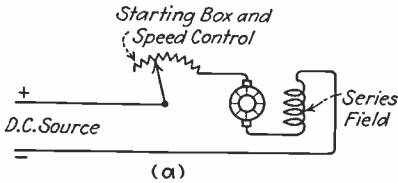


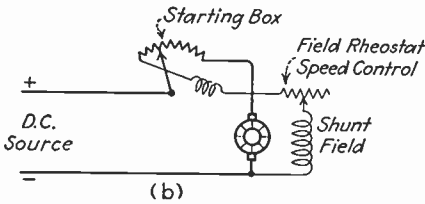
FIG. 386.—Field effects in motor.

connections shown in Fig. 387. It is to be remembered that the only difference between a direct-current motor and a direct-current generator is that one is driven mechanically and sets electricity in motion, while the other is driven electrically and produces mechanical energy.

a. *The Series Motor.*—This type of motor, shown in diagram, Fig. 387(a), is used on services requiring quick acceleration with a heavy load. The load must never be taken off a series motor because, if this happens, the motor will race and tear itself to pieces. The series motor is not used in radio service and, therefore, will not be further discussed here.



b. *The Shunt Motor.*—The shunt motor [Fig. 387(b)] is used for service in which constant speed is an important factor. This type of motor, once it has reached full speed, will maintain this speed in a very steady way, regardless of variation of the load.



In order to understand why this type of motor behaves in this way, study its connections. Assume that a current is impressed across plus and minus at the point marked D.C. SOURCE. This will excite the field windings. As long as the impressed e.m.f. remains constant, the field strength and excitation current will remain constant. A current which is in unit proportion to this excitation current will be forced through the armature. The motor armature now commences to rotate and attains a certain speed. The armature is now producing a counter e.m.f., due to its conductors cutting the magnetic field, which, when the motor is running at no

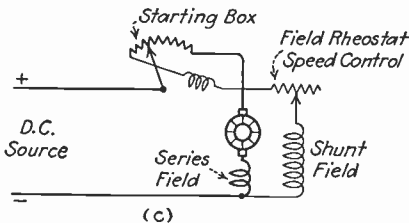


FIG. 387.—Types of direct-current motors.

e.m.f.; therefore, at no load, practically no current (amperes) is flowing through the motor.

Assume that a load is suddenly thrown on the motor; it slows down slightly but not very much. As this happens the armature will cut a smaller number of lines of force per second and will produce a smaller counter e.m.f. Because of this the current (amperes) flowing through the armature increases. This increases the torque (turning effort) of the armature and the original speed of the motor is maintained. Now, if the load is taken off, the motor speeds up somewhat, the armature cuts more lines of force per second, which in turn induces in the armature a greater counter e.m.f., preventing the motor from gaining more speed. It is, of course, true that a shunt motor does vary slightly in speed, but for correctly designed machines this variation is so slight that for the purpose of theoretical explanation it may be disregarded.

c. *The Compound Motor.*—This type of motor has the characteristics of both the series and the shunt machine. It is employed in such services as require a large starting torque and a constant speed under load. Because of its series field, it is able to gain speed quickly under a load; and because of its shunt field it maintains an even speed when the load is suddenly taken away or varied. These machines are

designed especially for constant speed under quick changes from no load to full load. In radio-telegraph transmitters the motor is practically running light until the key is depressed when full load is thrown upon it immediately. A circuit diagram of a long-shunt compound motor appears in Fig. 387(c). See Sec. 351*d* on Field Excitation for an explanation of long- and short-shunt compound machines which applies to motors as well as generators.

d. The Differentially Wound Motor.—The differentially wound motor is a compound machine on which the series field is connected so that it is in opposition to the shunt field. This results in an automatic regulation of speed under varying loads. Remember that a weak field permits the motor to speed up. Now look at the diagram (Fig. 388) which shows the fields as being opposed one to the other. Imagine the motor running at normal speed. If a sudden load is thrown on the machine it will slow down just slightly. As this happens the counter e.m.f., produced by the motor armature decreases, allowing more current to flow through the armature circuit. The series field of the differentially compounded motor is, as in all other compound

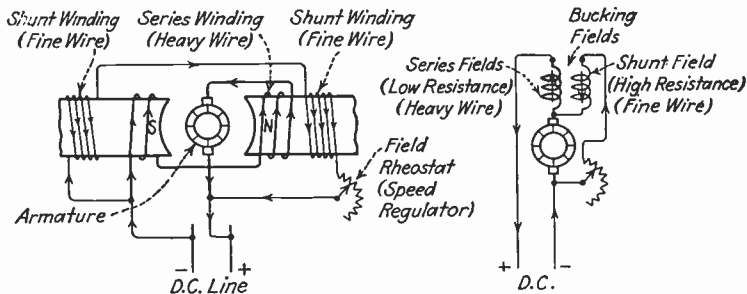


FIG. 388.—Differentially wound compound motor.

machines, in series with the armature and, therefore, if the armature current is increased, the series-field current increases, which in turn increases the field of force produced by this field. But this increase in field density, instead of augmenting the field of the whole machine, bucks the shunt field (it is connected in opposition), and the resultant is a weaker field which allows the machine to increase its speed.

When the load is decreased, the current in the armature is decreased, and the series or opposition field, as it might be called, is also decreased, which allows the shunt field to exert its full influence to produce a strong field and thereby prevents the armature from turning too rapidly.

It is seen, therefore, that the series field of a differentially compounded motor serves as a very good speed regulator. This type of motor is used to a considerable extent with radio equipments.

e. The Interpole Motor.—The construction of the interpole motor is identical with that of the interpole generator described in Sec. 346. The function of the interpole is there explained with the aid of diagrams.

f. Universal Alternating- and Direct-current Motors.—If the terminal connections to a series motor [Fig. 387(a)] are reversed, the direction of rotation of the machine will still be the same because the direction of rotation depends upon the relative direction between field and armature currents. It is easily seen that as long as the connection of the armature to the field has not been disturbed a reversal of terminal polarity will not affect direction of rotation. Such being true, an alternating current may be applied to the terminals of any series motor and the motor will run satisfactorily. The only difference between a series motor designed for direct-current

operation and one designed for alternating-current operation is that allowances must be made in core design for eddy currents set up during alternating-current operation. Eddy currents are, of course, not a problem in direct-current machines. Many small motors, especially those designed for household uses, are series machines designed for alternating- or direct-current operation.

Note on Direction of Rotation: A series motor will run in the opposite direction to what the same machine would run when used as a generator for given connections and relative direction of current flow in the circuits.

A shunt machine runs in the same direction whether used as a motor or a generator for given current direction and circuit connections.

360. Alternating-current Motors. *a. The Synchronous Motor.*—The synchronous motor is in effect an alternator, run as a motor. The field is supplied with direct current and the armature with alternating current; if the machine is gotten up to synchronous speed, it will continue to rotate and is then a synchronous motor. The synchronous motor has the advantage of running at an absolutely constant speed or

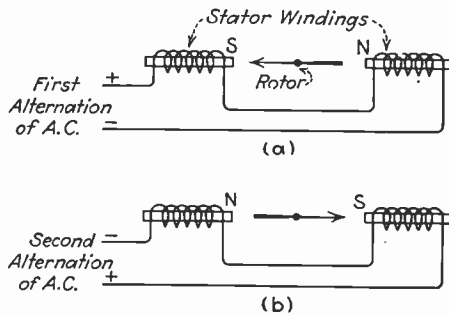


FIG. 389.—Principle of single-phase induction motor.

not running at all. If the load becomes too great, the motor simply falls out of phase and stops. It will burn up unless disconnected from the line immediately. Large synchronous motors are generally started by a smaller direct-current motor, attached to the same shaft, which becomes the exciter for the fields after the motor is up to speed.

Under normal operating conditions, the synchronous motor draws a leading current from the line and in this way may be used to offset the lagging current drawn by induction motors on the line. When so used it is called a *synchronous condenser*.

b. The Induction Motor.—The induction motor differs from the direct-current motor in that it depends for its rotation upon a rotating field, in addition to the motor action produced in the conductors of the rotating element. A fundamental idea of how this rotating field is produced may be had by studying Figs. 389 and 390. Figure 389 shows the idea applied to the single-phase machine. In Fig. 389(a) the arrow points toward the left, and in (b) it points to the right, which is the condition for the first and second alternation, respectively, the field having reversed itself. If now a rotor is arranged with momentum enough to carry around a quarter-revolution while the field is reversing itself the motor will continue to run; the motor, however, would not be self-starting.

If a three-phase induction motor is connected to a three-phase alternator, a revolving field will automatically be produced in the motor as the three-phases of alternating current produced by the alternator are generated. If the arrow in the motor diagram (Fig. 390) indicates the rotor of the motor, it will take the relative positions shown in the three diagrams for a partial revolution of the alternator as shown. Thus the

rotor of the motor will follow the alternator rotor around, showing that a revolving field has been produced.

The field winding of an induction motor is called the *stator*. In small single-phase machines the stator consists of two windings, a starting winding and a running winding. The action of these two windings in causing the single-phase motor to be self-starting will be discussed farther on.

The induction motor is often likened to a transformer, the stator being the primary and the rotor the secondary. In some ways the action between the stator and rotor of the induction motor is similar to the action between the primary and secondary of the transformer and for this reason the induction motor is said to have a primary (stator) and a secondary (rotor). In the induction motor there is no electrical connection between the stator and the rotor, and in the transformer there is no electrical

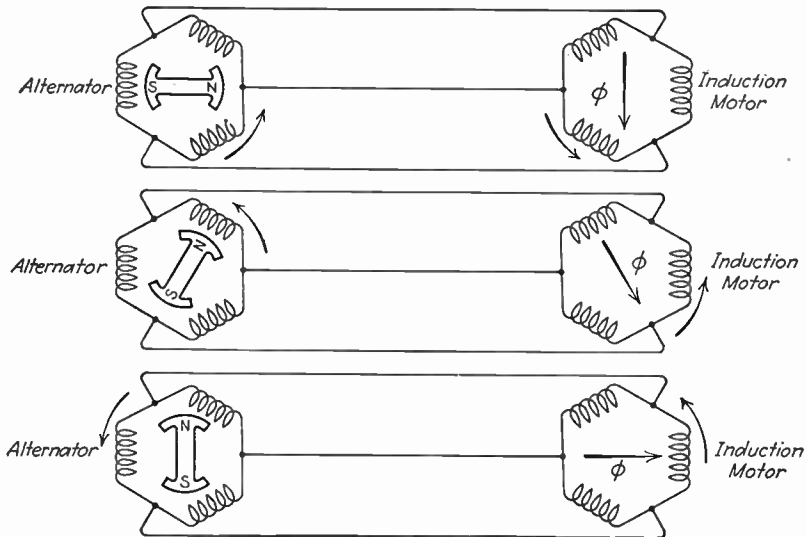


FIG. 390.—Principle of three-phase induction motor.

connection between the primary and the secondary; in this respect the induction motor and the transformer are alike.

The most commonly used secondary, or rotor, of the induction motor consists of a laminated iron core over which is fitted a *squirrel-cage winding* which is so called because of its resemblance to a squirrel cage. The squirrel-cage winding without the core is shown in Fig. 391.

How the Induction Motor Works.—In the assembled motor the squirrel-cage winding is placed in a running position between the stator coils of the machine, which coils produce a field surrounding the rotor. When the machine is running, the stator field is revolving and the rotor is revolving. The field revolves slightly faster than the rotor regardless of the operating speed of the machine. This causes the conductors in the rotor to be cut by the lines of force of the field. Now, whenever a conductor is cut by lines of force, it tends to move at right angles to the direction of the flux in accordance with the right-hand rule for motor action. All of the conductors are cut, and together they produce a torque on the rotor sufficient to cause the rotor to rotate and produce the rated power of the machine.

Therefore, the induction motor is caused to rotate by the action of the rotating field in tending to drag the iron core of the rotor around with it as it rotates, and by

the motor action of the conductors of the rotor as they are cut by the moving stator field.

All induction motors are not squirrel-cage machines; some have *wound* rotors with the terminals of the winding coming out to slip rings similar to the rotating-armature alternator. These slip rings may be connected to resistances for controlling the motor's speed.

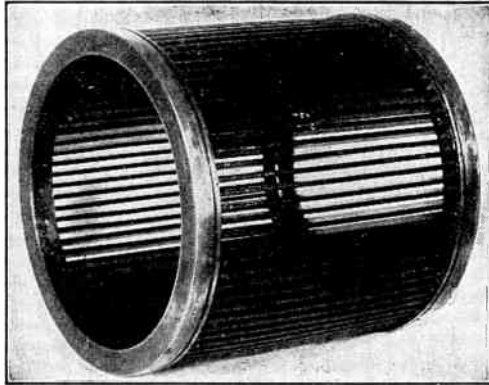


FIG. 391.—Squirrel-cage winding without core.

III. MOTOR STARTING, CONTROL, AND OPERATION

361. Hand-starting Boxes.—One of the most common types of hand starters is shown in Fig. 392. It is here shown connected to a shunt motor. The starting box is used on the larger motors, those of a size too big to be started directly from the line. The function of the starting

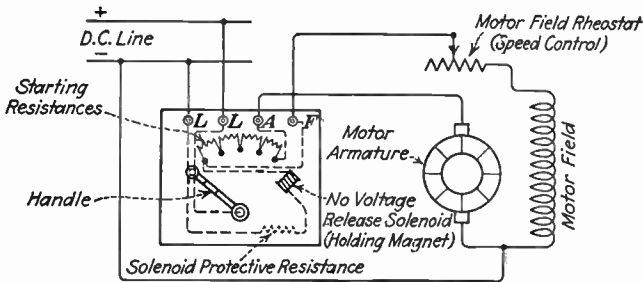


FIG. 392.—Hand starter, G.E. type.

box is to limit the armature current during starting and thus prevent the armature from burning out before it could build up the required counter e.m.f. to limit the current automatically as it does when up to speed. A study of the diagram will show that the starting box consists of a number of resistance coils which are gradually cut out of the circuit as the handle of the box is pulled over toward the right. Every hand starter has either a no-voltage-release or a no-field-release magnet. This magnet is in the field circuit, and should the current in this circuit

be interrupted for any reason it loses its magnetism and the handle flies back to the "off" position. The handle is fitted with a spring which tends to keep it in the off position, and it will not stay over to the *full-speed* point unless the attraction of the release magnet is strong enough to hold it there.

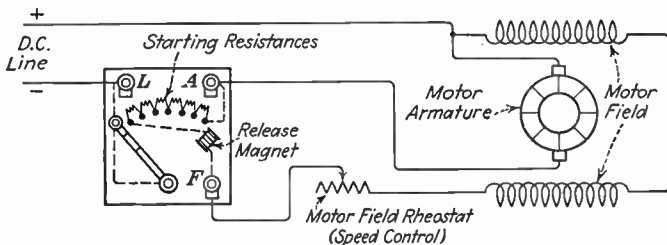


FIG. 393.—Hand starter, C-H type.

The hand starter shown in Fig. 392 is the type made by the General Electric Company and has four terminal posts, indicated on the diagram and on the box by the letters *L, L, A, F*, which mean *lines, armature, and field* respectively.

Another type of starter commonly used is made by the Cutler-Hammer Electric Manufacturing Company. This starter is fitted with three terminal posts and is connected in the circuit, as shown in Fig. 393, somewhat differently from the previously described four-point box.

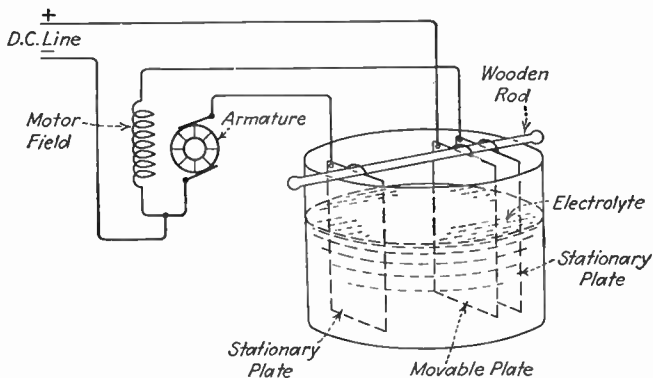


FIG. 394.—Water rheostat.

The handle of the starter must be pulled over quite slowly, as the motor must be started correctly, neither too slowly nor too fast. If the motor is started too quickly, the fuses will blow; or if the circuit is provided with circuit breakers, they will immediately open. The reason for this is that the motor armature, as was explained before, has a very low resistance and therefore draws a very heavy current until it attains its regular speed. There is no special rule to follow in starting a motor

with a hand starter, but the speed at which the handle should be moved over can soon be determined by gaging the acceleration of the motor. Radio motor generators used on shipboard are relatively small and are usually started in from 15 to 30 sec.

The resistance coils of the hand starter are meant for temporary starting duty only, and, if the current is allowed to flow through them for

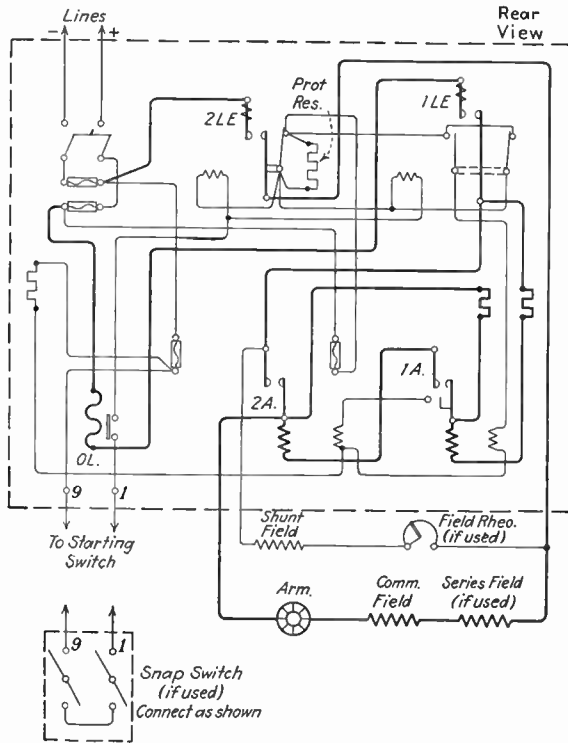


Fig. 395.—Magnetic-relay-type automatic starter (G.E.).

too long a time, it might result in one of them burning out. Pulling the starting handle over too slowly might, therefore, burn out the box. The remedy for a burned-out resistance coil is to short-circuit it until it can be repaired. If more than one resistance unit is burned out, however, it would very likely be very dangerous to use the box at all. It would be better if some emergency device, such as a water rheostat (Fig. 394), were used. Small motors of a fraction of a horsepower are generally connected directly to the line without a starter.

362. Automatic Starters for Direct-current Motors. a. Magnetic-relay Type (General Electric Company).—This automatic starter is shown in Fig. 395, which should be referred to in connection with the following explanation of its operation. If the line switch is closed and the remote-control starting switch is open, all contactors in the automatic starter will be open and no current flows. As soon as the

starting switch is closed, current flows through the solenoids of the relay, 1-LE and 2-LE, which immediately close, allowing current to flow through the armature and field of the motor. This current is limited by the two series resistors in this circuit. All heavy lines on the diagram are those carrying armature current, while the light lines carry field and control current. Current flowing through the armature also flows through relay 1-A. However, this relay is provided with an additional winding through which current is now also flowing. The polarity of the magnetism induced in the core by these two windings is opposite, so the resultant magnetism is very nearly zero; hence this relay does not close immediately. However, as the motor builds up speed and the back e.m.f. increases, the current through the armature becomes less. Eventually, the magnetism in this relay, due to the armature current, becomes so small that the field due to the auxiliary coil predominates and the relay closes. When the relay closes, its contacts close two other circuits, both going to the relay 2-A. The main armature current now goes through the main solenoid winding on relay 2-A, which is a low resistance, instead of going through the series resistor. Of course, there is another resistance still in the circuit. The other contact of relay 1-A closes the circuit to the auxiliary solenoid winding on relay 2-A. The current through the armature is therefore increased, since some of the series resistance has been cut out of the circuit. The operation of relay 2-A is similar to the operation of relay 1-A, that is, this relay does not close at first, but, if the back e.m.f. of the motor builds up to a certain value, this relay (2-A) closes, and the line current goes straight into the armature without any resistance in series. Item OL is an overload relay, which stops the motor-generator set, if the armature current increases beyond a safe value. This overload relay must be reset by hand after opening. The extra contacts on relay 2-LE cut a resistance into the circuit in series with the loading solenoids of relays 2-LE and 1-LE.

When supplied to start a radio-transmitter motor generator, this automatic starter, in addition to starting and stopping the motor, also opens the field circuit of the high-voltage generator by closing an auxiliary contactor usually located in the transmitter unit.

b. Dashpot Type.—This type of automatic starter is so called because the acceleration of the starting arm is controlled by a dashpot filled with oil. A circuit diagram of this starter as used on Mackay radio transmitters is shown in Fig. 396.

The dashpot-type starter is of the time-limit class of starters with four accelerating steps. It is provided with a thermal-overload relay which shuts down the motor generator in case of overload. If this relay opens, it must be reset by pressing a button on the front of the case before the motor generator can be restarted.

The motor starter is controlled from the operating table by a three-wire "start-stop" push-button. This type of push-button makes it necessary to restart the motor generator, by pressing the starting button, whenever the supply line is opened.

The operation of the starter may be understood by referring to the wiring diagram and studying the following explanation.

When the starting button is pressed, the solenoid of the starter is energized and the plunger starts to close five contact arms above it. Owing to an oil dashpot beneath the solenoid the plunger cannot rise immediately and from 5 to 10 sec. elapse before all of the contact arms close. The first arm to make contact is the one on the right which keeps the solenoid energized when the "start" button is released. The arm on the left then closes and completes the supply circuit to the motor through a suitable starting resistance mounted on the starter box. The motor will then come up to speed. As the contact arms continue to close and short-circuit sections of the starting resistor, the speed of the motor will gradually increase and will reach its final value after the last contact has closed. At the end of the plunger's travel,

contacts at the bottom of the right-hand arm will close, and the upper contacts on the same arm will open. This connects two small resistors in series with the solenoid, which reduces the current in the solenoid while the motor generator is running.

363. Starting Induction Motors.—The single-phase induction motor may be started by hand, which is generally impractical, or it may be started by using a phase-splitting circuit which consists of two stator windings, a starting winding, and a running winding.

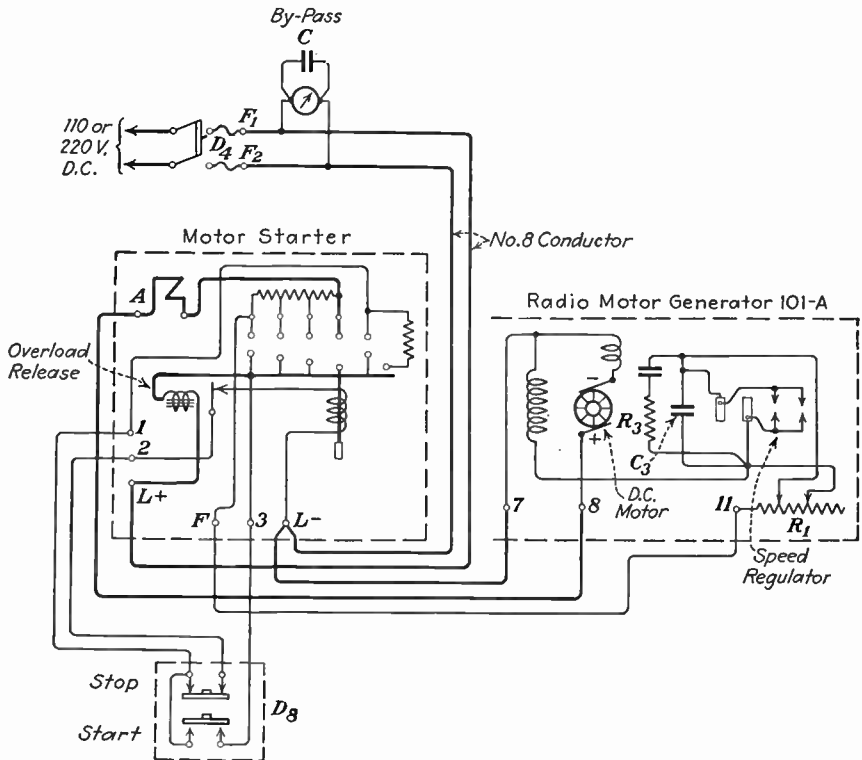


FIG. 396.—Dashpot-type automatic starter.

Explanation: In the phase-splitting circuit the current is sent through the stator by two paths in parallel (the starting and running windings). The inductance of one path is considerably more than the inductance of the other path which causes the current in the higher inductance path to lag behind the current in the other winding. This lag in the starting winding is 90 degrees behind that of the running winding, which causes a revolving field which starts the motor. After the motor is up to speed, a centrifugal cut-out cuts the starting winding out of the circuit and the motor continues to run with only the running winding in operation. The starting winding consists of relatively few turns of fine wire; the running winding consists of many more turns of heavier wire. In some motors an inductance coil in the base of the machine is connected in series with the starting winding to give it the necessary added inductance.

Small three-phase motors (squirrel cage) are generally connected directly to the line through the special starting switch, shown in Fig. 397. Note that in the starting position heavy fuses are provided for the heavy current drawn from the line when starting, and in the running position smaller fuses are used to give the necessary protection over long periods.

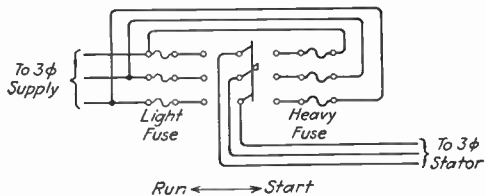


FIG. 397.—Starting-switch circuit for small three-phase motors.

One of the most widely used starters for two- and three-phase induction motors is the *induction starter*, shown in Fig. 398. This type of starter utilizes the autotransformer together with a switching device which connects the stator across the autotransformer in the *starting* position and across the lines in *running* position. A tracing out of the current paths in the schematic diagram will show how this is accomplished. The long narrow rectangles shown above and below the contact points on the switch are the movable contacts of the switch.

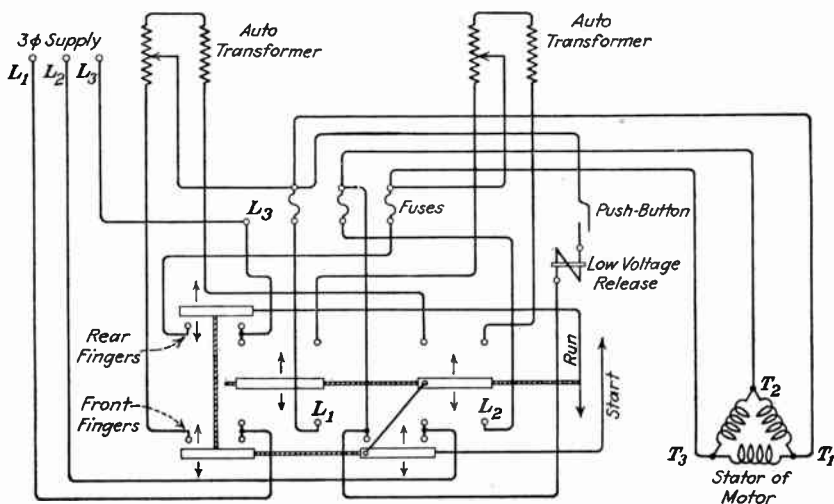


FIG. 398.—Induction starter for large induction motors.

For large machines these starters are arranged to apply the line voltage in three or more steps while for lighter machines the voltage is applied in two steps.

A feature of this switching arrangement is that it disconnects the transformer windings from the line entirely during running position.

In the starting position, in the starter shown, the fuses are not in the circuit; they are thrown in the circuit when the switch is thrown to running position. In some induction starters heavy fuses are cut into the circuit for starting and lighter fuses for running. These starters are usually fitted with no-voltage release devices and sometimes with overload circuit breakers which take the place of fuses.

Another method of starting an induction motor is to provide the machine with a commutator and brushes as a direct-current machine, except that the rotor is not connected to the supply line. This type of machine is known as the *repulsion motor*. The brushes are short-circuited, and, when the current flows in the field, a heavy current is induced in the rotor windings producing a strong starting torque. When the motor is up to speed the brushes are lifted from the commutator and the machine runs as a simple induction motor.

364. Manual Speed Control.—*a. The speed of direct-current motors is regulated by varying the field strength of the machine. This is done by connecting a field rheostat in the field circuit to regulate the field current as shown in the motor circuits of Fig. 387. Notice that in the compound machine the field rheostat is connected in the shunt-field circuit.*

Some motor field rheostats are provided with an automatic release which consists of a magnetic device, the solenoid of which is connected across the armature terminals, which short-circuits the rheostat resistance when the motor stops, thus providing a strong field when the machine is started. When the machine gets up to a predetermined speed, this automatic switch opens and puts the field rheostat in the field circuit.

Generally the resistance of the field rheostat for shunt and compound machines is about two-thirds of the field-resistance value of the machine. This allows an increase in speed of about 25 per cent. If greater speed variation is desired the resistance of the field rheostat may be very much greater than two-thirds that of the field.

b. The speed of the single-phase induction motor is variable over a small range and may be changed by varying the voltage applied to the motor, the torque being proportional to the square of the voltage for any given slip.

The speed of two- and three-phase induction motors varies with the load, much like the shunt motor. Under no-load conditions the speed of the motor is very nearly the speed of the field which is called the *synchronous speed*; and where the synchronous speed of an induction motor is specified, it usually means the no-load speed. The synchronous speed of two- and three-phase machines may be calculated by the following equation:

$$\text{Synchronous speed} = \frac{120 \times \text{frequency}}{\text{number of poles}} \quad (179)$$

The difference between the speed of revolution of the stator field and the rotor of the induction motor is called the *slip* of the machine and is calculated by the following equation,

$$\text{Percentage of slip} = \frac{\text{synchronous speed} - \text{actual speed}}{\text{synchronous speed}} \times 100 \quad (180)$$

There are two common methods of controlling the speed of an induction motor: (1) *by varying the voltage* applied to the stator, using variable-resistance controllers, or by the more complicated compensator method by which coils are thrown in and out of the circuit thus varying the applied voltage; or (2) *by using a wound rotor* with slip rings and connecting a variable resistance between these slip rings.

365. The Centrifugal Speed Regulator.—This regulator which is shown attached to the motor shaft in Fig. 396 is capable of holding the speed of the motor constant to within 2 per cent for any variation in line voltage normally encountered on board ship. By maintaining the motor speed constant, the generator voltages can be held at constant values.

The centrifugal speed regulator consists of a pair of contacts mounted on the end of the motor shaft. One contact is rigidly attached to the shaft, while the other is carried on a spring-supporting arm. The contacts are connected through slip rings to the ends of the fixed resistor R_1 which is in series with the field of the motor. When the motor is rotating close to rated speed, the arm vibrates so that the contacts touch and short circuit R_1 periodically. If the motor slows down due to low line voltage, contacts either do not make contact or make contact infrequently. This causes R_1 to reduce the field current in the motor, and makes it speed up. If the motor speeds up too much, the contacts touch frequently or remain in contact. This increases the motor-field current and slows down the motor. The motor will, therefore, run at a constant speed determined by the average effect of the making and breaking of the contacts. The speed will be almost independent of small changes in line voltage or load. To eliminate excessive sparking at the contacts, the condenser C_3 is connected across the terminals of the slip rings.

366. Operation of Motors. *a. General Requirements.*—As the series motor is not used in radio work this discussion will be confined to shunt and compound direct-current motors and to the induction motor.

The first consideration in the operation of motors is to make sure that all connections are properly made. There are ample circuit diagrams in this book to show how all types of motors should be connected. Improper connections may easily result in serious and costly damage to the machine and associated apparatus. Check up on the starting box to see that the A terminal is connected to one armature terminal of the motor; that the F terminal on the box is connected through the field rheostat to one field terminal of the motor; and that the L terminal or terminals on the box are connected to the line. On a four-post box the

remaining unconnected terminals of the field and armature are connected together and then in turn connected to the proper line terminal on the box as indicated in the wiring diagrams. Check to see that the field rheostat is connected properly.

When using a hand starter, care must be taken not to pull the handle over too fast or too slow. If it is pulled over too fast, severe sparking will occur at the brushes, the overload circuit breaker might trip, the fuses blow, or the motor itself might be damaged by an excessive current flow in the armature which has not yet had time to build up a current-reducing counter e.m.f. If it is pulled over too slowly, the resistance units in the starting box might burn out as they are designed only for temporary duty. The average motor will start in from 15 to 30 sec., but a good general rule to follow is to observe the amount of sparking at the brushes to see that it is not too severe and to listen to the general acceleration hum which should be a constantly rising tone until the motor is up to full speed. Severe sparking at the commutator when starting may mean that the starting-box handle is being pulled over too fast; try pulling it over slower.

Special care must be taken not to open the circuit of a shunt field as, if this field is removed, the armature will not produce a counter e.m.f.; the motor will then race itself to pieces or burn itself out. If by accident the field circuit is opened, the motor should be immediately stopped.

To stop a motor, open the main-line switch which will remove the power from the machine and cause it to stop.

b. The compound motor has a shunt field of many turns of fine wire and a series field of fewer turns of heavy wire. The series field should be connected in series with the armature and the remaining terminals considered as the armature terminals. The motor may then be connected to a starting box as if it were a shunt motor.

In connecting the series field in series with the armature, it must first be decided whether the motor is to be cumulatively operated or differentially operated. If cumulatively connected, the series field should be connected so that the current flows through it and the shunt field in the same direction. If differentially connected, the current must flow through the series and shunt fields in opposite directions.

c. To reverse the direction of rotation of any type of direct-current motor, it is only necessary to reverse the current flow in either the armature or field of the machine, but not in both. Practically, this may be easily done on a shunt motor by disconnecting the armature leads and reversing them, or, if the motor is fitted with a terminal block, by reversing the external wires connecting to the field posts.

On a compound motor more care is necessary when making reversing connections because of the fact that the machine has two fields, a series field and a shunt field. To reverse a compound motor, (1) reverse the

armature connections alone or (2) reverse the current in both the series and shunt fields while leaving the armature-current flow as it was.

With the above requirements in mind it is a simple matter to connect a reversing switch in the motor circuit if necessary or desired.

IV. MOTOR GENERATORS AND POWER-CONTROL APPARATUS

367. Advantages and Disadvantages of Motor Generator.—The advantages of the motor generator over other sources of power supply lie in the easy voltage control and good regulation of output which is obtainable at full-rated voltage. The motor generator is an exceptionally rugged machine and may be located at a point remote from the transmitter and remotely controlled. It may also be operated from either alternating- or direct-current lines by simply adapting the driving motor to the power available.

The disadvantages of the motor generator may be summed up as follows: Poor regulation at less than rated voltage; may require a long time to repair if breakdown occurs; may be noisy and cause vibration; requires constant attention to lubrication; and entails a high installation cost. It may be necessary to run high-voltage leads within reach of operating personnel, which may be dangerous to life.

368. The Dynamotor combines a motor and generator in a single-frame machine which requires but two bearings. The single armature of the machine consists of two separate windings which come out to two commutators, or two pairs of slip rings, or a commutator and slip rings, depending upon the use to which the machine is to be put. If the machine is to run on direct current and deliver alternating current, it would have a commutator on the motor end and a pair of slip rings on the generator end. If the machine is to run on direct current of one voltage, and generate direct current of another voltage, it would have two commutators. It is, therefore, easy to see what a machine would need in the way of a commutator or slip rings by simply considering the conditions in the circuit.

The dynamotor has but one field winding, which serves to excite both the motor and the generator windings of the armature.

Owing to the fact that the armature reaction of the motor and generator windings are equal they balance one another out, and the machine has no armature reaction. The total losses in the machine are, therefore, less than the losses in a motor-generator set, raising the efficiency of the machine proportionately. Because the field winding does not have to overcome armature reaction, a lighter field coil may be used and the machine may, therefore, be smaller and weigh less than a motor generator.

The disadvantage of the dynamotor is that the output voltage cannot be varied without varying the input or motor excitation. It is, there-

fore, impossible to get the smooth regulation of voltage with a dynamotor that is possible with the motor generator.

The dynamotor is finding extensive use on aircraft and automobile installations for supplying plate power, thus eliminating the need for heavy plate batteries.

369. The converter, which is also known as a *rotary* converter or a *synchronous* converter, performs the same function as the dynamotor and also has the advantage of combining the functions of a motor and generator in a single-frame machine with a single armature. The converter has but one field, the same as the dynamotor, and also one armature winding, wherein it differs from the dynamotor. The armature is fitted with a commutator on one end and slip rings on the other, and the machine is used to convert alternating current to direct current or vice versa. When it is used to convert direct current to alternating current it is called an *inverted converter*.

The disadvantage of the converter is the impossibility of getting smooth voltage control because the motor and generator functions are performed by a single set of windings, which means that the input and output characteristics cannot be controlled independently.

On the other hand, where voltage control is of secondary importance both the dynamotor and converter have the advantage of costing less than a motor generator, and also of taking up less space and of being lighter than a motor generator of similar rating.

370. Double-current Generators.—If a dynamotor or converter is driven by a suitable prime mover, such as a gasoline engine or fan drive, two kinds of voltage can be taken from the machine. This principle is made use of on aircraft installations where such *double-current generators* are used to supply *A*, *B*, and *C* voltages to the transmitter and receiver without the use of storage batteries as the primary source of power. In order to get a constant voltage with the variation in armature speed which is likely with a fan drive, the propeller is of the self-regulating type which automatically varies the pitch of the blade according to the wind pressure. This maintains a constant speed of rotation and makes an electrical voltage regulator unnecessary. These wind-driven generators have also been fitted with differential fields, in addition to their shunt fields, which act as automatic electrical voltage regulators.

371. General Maintenance.—Under general operating conditions the only care required by dynamo-electric machinery is that it be kept properly lubricated and reasonably clean with well-fitting brushes and tight connections. On direct-current machines special attention must be given to the commutator to avoid sparking at the brushes with consequent harmful results. No motor or generator should be run overloaded for long periods. Most machines are designed to withstand overloads for a short period without damage to the machine.

The commutator on a direct-current machine requires more attention than any other part of the machine. Bad operation of the machine makes itself immediately apparent on the commutator as it is this part of the machine which gets the most mechanical and electrical wear. A properly functioning commutator should have a deep chocolate color, especially when carbon brushes are used, and should acquire a smooth polished surface. Roughness on the commutator may be detected by feeling the surface of the bars with the finger nails. On low-voltage machines this may be done while the machine is in operation, but on high-voltage machines the power should be entirely removed from the machine and the armature turned over by hand. Where there is any doubt as to the safety of touching the commutator, while the machine is in operation, the latter method should always be followed.

If the commutator has any roughness, it should be smoothed down with fine sandpaper. Never use emery cloth, as emery dust is metallic and particles of this dust may serve to short-circuit the commutator bars. Number 00 sandpaper is good for this purpose with a little oil applied to give a polish.

A very rough or badly grooved commutator will need to be turned down in a lathe to a new level of smoothness or, if the commutator is in very bad shape, a new commutator may have to be installed. These operations are generally beyond the scope of the radio personnel, this work being best performed by an expert repair man.

A commutator may with wear become slightly off center, which condition will cause the brushes to move up and down as the commutator revolves. To eliminate this it is necessary to turn the commutator down in a lathe until it runs on a true center. Before blaming the commutator, however, make sure that the motor or generator shaft has not been sprung so that the entire armature is off center.

372. Brush Troubles.—Most common brush troubles may be caused by (a) too low pressure on the brush, (b) brushes “off” electrical neutral, (c) brush not fitting properly, (d) wrong kind of brushes used.

a. Too low brush pressure will cause a slight arc to be produced between the brush and the commutator resulting in a roughening and pitting which is aggravated as the condition continues. Set the brush in the holder with enough pressure so that a good contact results when the machine is operating.

b. It should be possible to shift the brushes to a point where little or no sparking appears on full load. This is the electrical neutral and the proper operating point for the brushes. Do not attempt to shift to an electrical neutral on an interpole machine as the interpoles are designed to eliminate sparking and this should be unnecessary. On interpole machines, sparking at the brushes is caused by some other cause than the brushes being off the electrical neutral.

c. As brushes sometimes differ slightly in size, it sometimes happens that a brush does not fit into the holder properly. This may not be noticed until improper operation of the machine is traced to this cause. The remedy is obvious; the brush should be of the right dimensions to fit properly into the holder. The holder mechanism

itself must keep a proper pressure on the brush during operation allowing for the wearing down of the brush as operation continues.

d. Many times, when a machine is shifted from one service to another where the load conditions are different a different type of brush will be necessary. If a trial of various types of brush available does not result in proper operation of the machine, the manufacturer should be consulted for advice on the proper brush to use.

373. Bearings.—Wherever there is a moving part in a machine there is friction; wherever there is friction there is wear and heat; and wherever there is wear there should be a lubricant. If a piece of steel is coated with oil, a thin film forms over the surface of the steel which cannot be removed by friction even though the steel be vigorously rubbed with a dry cloth. It is this film of oil which will wear out eventually to a point where it will be ineffective. Friction will then result between the surfaces of the rotating steel shaft and the machine bearing. It is the purpose of proper lubrication to keep the film of oil thick enough on the surfaces subject to frictional wear to prevent the steel surfaces of the shaft from being worn at all, and to keep the wear on the bearings down to a minimum.

If from improper lubrication or other causes a bearing on a machine gets unduly warm or hot, a liberal supply of cool clean oil should be applied immediately. If possible the machine should be slowed down but not stopped and the oil flushed in the bearing as much as possible. The machine should be kept running to prevent a contraction of the bearings around the rotor shaft which is commonly referred to as *freezing* and which causes the shaft to be jammed in the bearing. Under these conditions the shaft cannot rotate until pried loose from the bearing which may be an extensive job. In cases of emergency, applications of ice held in ice-water-soaked towels may be applied to the outside of the bearing. If the machine is fitted with an adjustable bearing cap, this should be slightly loosened. Graphite mixed in with oil is beneficial if applied to a hot bearing.

In every case of overheating, the cause should be ascertained and steps taken to prevent a recurrence. All oil or grease used for lubrication purposes should, of course, be free from foreign matter and suited to the service to which it is applied.

When a machine is fitted with oil rings, these should be inspected to see that they operate properly, being loose and dipping freely into the oil well. If the machine is fitted with grease cups, give each cup a quarter-turn once each week. If a machine is fitted with ball bearings, these should be cared for according to the following directions, to apply particularly to the motor-generator set shown in Fig. 371, but which apply generally to all ball-bearing machines.

*The Care of Ball Bearings*¹—*Lubrication.*—As motor generators cannot always be placed on shipboard with the shaft fore and aft, there may be times when they

¹ Directions issued by Crocker Wheeler Company.

will be subjected to severe rolling and it is, therefore, desirable in general to use a medium grade of grease similar to No. 3 Keystone grease¹ which is made by the Keystone Lubricating Company of Philadelphia, Pa. A slightly lighter grease could be used in cold weather or where there would not be much rolling. Any similar good grade of grease would be satisfactory but it must be absolutely neutral or test free from acid or alkali.

The other important requirement is that the grease must contain no fibrous material and must withstand a temperature of 100 degrees C or higher without any material change in its consistency when cold. Some greases, upon being heated, separate into a thin oil and a relatively viscous soapy material. This kind of a grease should not be used unless absolutely necessary. If the bearings are filled full of grease it will be found that a portion of this will, after running for a time, work out of the ends of the bearings, leaving, however, sufficient to properly lubricate same. After operating for a month or so it is desirable to thoroughly flush out the bearings with gasoline or kerosene oil and fill with a good grease. After this, if the bearings are kept free from all dirt, grit, etc., they should not need cleaning and refilling with grease more often than every six months.

Cleanliness.—This is an extremely important matter with ball bearings and, therefore, all foreign material except the grease must be carefully excluded. Care must be taken that no dirt or grit of any kind is carried into the bearing with the new grease, or otherwise, when, for any reason, the bearing caps are removed.

When the machine is not equipped with ball bearings, the oil wells and rings should be inspected at least once a month or oftener to see that they are well lubricated.

Bearings in the smaller machines are sometimes of the roller-bearing type, while in the larger machines the oil-ring type of self-oiler is commonly employed with babbitt-metal or bronze bearings.

374. Trouble Shooting.—Some of the most common troubles encountered in the operation of dynamo-electric machinery are open circuits and short circuits in the armature and field windings, field rheostats, or starting resistances. In connection with spark transmitters on ship-board, a grounded protective condenser may be the cause of trouble. It is our purpose now to describe in detail tests to locate these troubles.

a. Open Circuit in Armature.—An open circuit in an armature coil will manifest itself in severe sparking at the commutator every time the open-circuited coil passes under a brush. An armature may be tested for an open coil by using a test lamp as shown in Fig. 399(a). When the open coil is reached, the lamp will burn dimly or it may go out, depending on the resistance of the armature winding. The machine is entirely disconnected from the line when this test is made.

¹ Texas Oil Co.'s grease, Marfak No. 3, or equivalent, is recommended by the R.M.C.A. for their ball-bearing motor generators.

Sometimes an armature coil will have a break in the wire which will only pull apart when the machine is rotating, owing to the centrifugal force then present and acting on the wires. Under these conditions the break may still be present but making electrical contact when the machine is at rest at which time it will not be apparent when the armature is tested as shown in Fig. 399(a). This is known as a *flying open* circuit, and its detection is a job for an expert with ample testing equipment.

b. *Short Circuit in Armature.*—A short-circuited armature coil may be suspected in a motor, if, when starting the motor, the armature does not turn over until three or four points on the hand starter have been passed

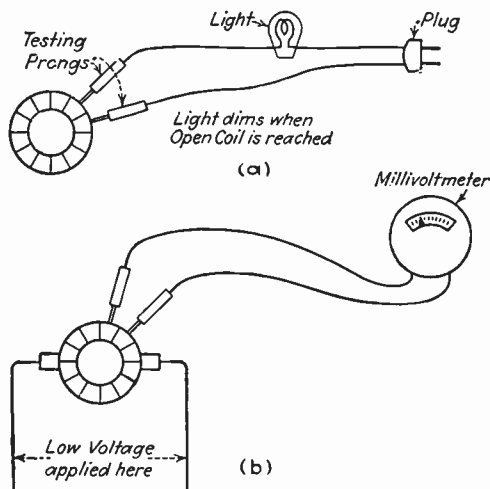


FIG. 399.—Method of testing armature: (a) for open circuit; (b) for short circuit.

(or with an automatic starter, delayed starting), then runs very unsteadily at the low starting speeds, and finally heats badly in one place on the armature when the motor has been running a short time. An easy test for a short-circuited armature coil is to apply a low voltage (from a battery or through a resistance or lamp bank from the line) to the brushes, as shown in Fig. 399(b), and then testing by the potential-drop method across each coil using a millivolt meter as shown. When the test prongs have made connection with the short-circuited coil, the reading on the meter will drop to very nearly zero, or certainly to a much lower value than the readings across the good coils. The armature should be disconnected from the rest of the motor when making this test to prevent the current from taking any other path than through the armature coils.

c. *Open or Wrong-polarity Field Coils.*—To test the polarity of adjacent field coils, pass current through the windings and place two

nails, one on each pole face pointing toward each other. If the polarity is right, they will attract each other; if it is wrong, they will repel one another. On some machines, where the residual magnetism alone is not enough to hold the nail to the pole face, a dead coil may be detected in this way.

Another test is to place a lighted carbon-filament lamp between the pole faces and watch the effect on the filament loops. The filament will be drawn together or forced apart depending on the direction of the flux. If a field coil is dead, no effect will be produced on the filament of the lamp.

d. Grounded Protective Condenser.—What is known as a grounded protective condenser, such as is used on spark transmitters on ship-board, is a broken-down condenser resulting in one plate of the condenser being electrically connected to the center post which in turn is connected to the ground. With such a punctured or grounded protective condenser one side of the line is grounded, which may result in damage to the machine if not corrected. A damaged protective condenser should be replaced, or cut out of the circuit entirely.

375. Protective and Control Devices. *a. Fuses* are placed in a circuit to protect the circuit and the apparatus connected to it from the damaging consequences of short circuits or other forms of overloads and to prevent the possibility of overheated electrical circuits and apparatus causing fire to material with which it is in contact or to which it is in close proximity.

Fuses are made of quick-melting alloys, the reason being that, if the circuit is overloaded, the current will rise and the consequent heat will quickly melt the fuse thereby opening the circuit and preventing damage. To prevent damage or fire from a blown fuse, the fuse strip is usually enclosed in some sort of a fireproof and insulating material. The fuse is such a common appliance that a detailed discussion of the construction of the various types will not be entered into. Suffice it to say that fuses are made to be renewable and non-renewable; the most common types are the cartridge fuse and the plug fuse.

When a fuse has blown, it should not be replaced until the cause of the blow-out has been ascertained and removed. Care should be taken that the same-size fuse is used in the replacement, for obvious reasons. When replacing a fuse it is safest to stand on insulating material; on shipboard avoid iron decks and do not hold on to iron bulkheads with one hand while changing the fuse with the other hand. Use insulated tools whenever possible and be very careful not to cause other short circuits on the switchboard or panel box while making the change.

b. Switches must be kept clean and in good condition mechanically, guarding against loose joints and faulty clips. Switches carrying a heavy current should be opened quickly to avoid blistering and burning

the contact edges. If such rough spots appear, they should be removed and the surface smoothed down with a file or emery paper. Switch blades should fit snugly into the clips to insure good contact and to avoid heating. A little vaseline placed on the switch blade occasionally makes the switch work easily.

c. *The Overload Circuit Breaker.*—In power circuits where the apparatus is to be protected from overload, fuses and overload circuit breakers must be used. On all panels in commercial equipment the overload circuit breaker is usually connected in the motor armature and generator

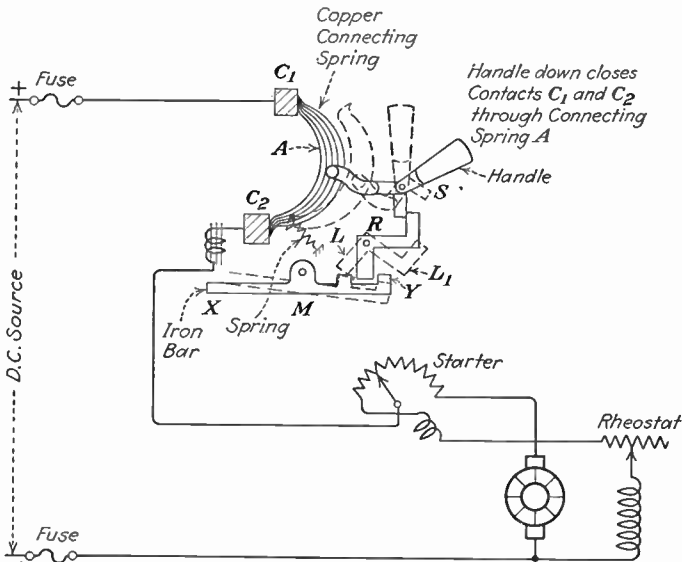


FIG. 400.—Details of circuit-breaker construction and circuit.

circuits. For example, if the plunger in the automatic starter should rise too quickly, or the hand starter be pulled over too quickly, the starting resistances would be rapidly cut out and a heavy current flow would flow through the armature winding. If this were not amply protected by fuses and circuit breakers, the armature winding, due to its low resistance, would be subjected to a heavy current and the result might be a burning out of one or more of the armature coils. The overload circuit breaker in Fig. 400 is illustrated as being connected in series with the armature circuit to protect against excessive current flow through the armature winding.

When the handle of the overload breaker is pressed down, a laminated copper spring *A* is pressed against contacts *C*₁ and *C*₂. This allows the current to flow through the armature and the overload magnet. It is quite obvious that as soon as a current flows through *A* the overload magnet is magnetized. The degree of magnetization depends upon the

amount of current flowing through the armature. This magnet will attract the iron bar M upward toward the magnet at point X . This results in point Y dropping, which causes the release of the latch L .

Latch L then takes the position of the dotted line L_1 , which causes a release of the handle from points R to S , incidentally releasing the copper spring A from contacts C_1 and C_2 , causing a break in the armature circuit.

The circuit breaker must be carefully adjusted so that it will not *trip* when the proper amount of current is flowing through the armature. This is accomplished by lowering the iron bar so that the distance between the bar and the magnet is increased to the proper distance. For example, if a motor, on starting, has been designed to safely draw 10 amp., then the iron bar must be moved far enough away from the overload magnet so that it will not break the circuit at that particular current flow. This is usually determined by experiment, although in many cases the amount of space for various amperage values is marked above the adjusting screw.

d. Vacuum Contact.—This type of contact gets its name from its construction which is shown in Fig. 401. Referring to this illustration it will be noted that the vacuum contact depends for its operation on the elasticity of a glass bellows in causing the mechanical actuation of contacts sealed in a vacuum. The glass bellows B , because of its shape and special temper, is highly elastic. A slight movement of the stem C is communicated to the movable block F , causing it to separate from the contact block G which is stationary. The spring E causes positive connection between contacts when no pressure is being applied on the stem. The contacts are contained in an evacuated glass tube A . Connections are made through the leads I .

The vacuum contact is free from the drawbacks of exposed electrical contacts. There is no exposed spark and the device is therefore flame-proof. Because of their operation in a vacuum the contacts produce a minimum spark or arc and this spark or arc can be further reduced by a shunted condenser. The contact members are free from corrosion, pitting, and sticking, when the contact is properly installed, which fact assures an almost infinite life.

A very slight movement (about 0.02 in.) at the end of the stem C will operate the contacts. Certain makes of vacuum contacts can handle up to 1,300 watts at 220 volts as fast as 40 breaks per second, the makes and breaks being clean, positive, and reliable.

When operating at the load specified above, the vacuum contact may be shunted by a 400-volt condenser. For non-inductive loads, the condenser can be of 0.0025-mf. capacity, and for inductive loads, 0.005 mf. when operating on direct current. For alternating-current applications, a 0.01-mf. capacity is recommended.

Various methods of using the vacuum contact are also shown in the illustration (Fig. 401 b and c).

376. Generator-ripple Reduction.—Direct-current generators designed to supply plate and bias voltages to vacuum tubes must be especially designed to minimize the minute variations in their output which in the aggregate are termed the *ripple* in the generated current. The armature of the machine must be carefully designed, especially the core which must have as many slots as possible.

As an aid in the elimination of brush sparking as a cause of ripple, it is necessary that the brushes be properly adjusted in the right position on the commutator so that the coils undergoing commutation are in the

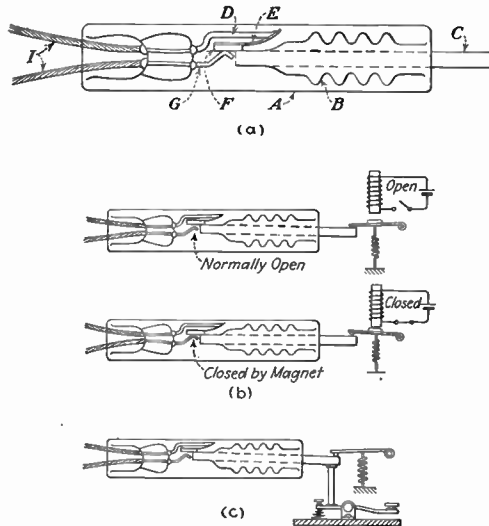


FIG. 401.—Vacuum contact. (Courtesy of Burgess Battery Company.)

neutral zone of the field. The correct position of the brushes is usually indicated, and care should be taken to see that they are placed in the proper position. The proper position is often not exactly between the pole pieces on account of distortion of the field under load, due to armature reaction.

It is also very important that the brushes used be of suitable material. Many different materials are used, and care must be taken to select brushes with sufficient lubricating qualities and which are made of material that will not wear down the commutator while making proper contact. Only brushes furnished by the manufacturer for the particular machine in question should be used. Much damage may be done by the use of brushes which are not suitable.

In larger generators, and in circuits in which the load is variable, as in keying the transmitter, it is desirable to use distributed-field windings, with compensating-field windings, to completely compensate for the

effect of the armature reaction on the field. This enables the machine to operate with perfect commutation under all conditions of load, and with sudden changes in load. Also a generator with compensating field windings may be short-circuited for short periods without damage, a strain which no other type of generator will withstand. In large machines for severe operating conditions, it is sometimes necessary to use both compensating windings and commutating poles to reduce brush sparking.

As mentioned before, it is the practice in some of the larger broadcasting stations to float a storage-battery bank of a voltage equal to that delivered by the generator across the output of the generator (the line) to absorb variations in current output and reduce ripple. An extremely pure direct current can be obtained in this way which is especially useful on circuits supplying power to speech-input equipment where the slightest ripple is amplified millions of times by the time it is emitted from the transmitting antenna. Information on floating batteries on the line for this purpose is contained in the following chapter, Sec. 399.

It is the more usual practice in most stations to use a ripple filter consisting of inductance in series and capacity in parallel with the line.

377. Generator-ripple filters¹ must filter out three kinds of ripple: (1) commutator ripple, (2) slot ripple, and (3) brush sparking.

(1) *Commutator ripple* is caused by uneven voltage distribution between adjacent commutator bars, which causes a spark to take place every time a brush leaves one bar and makes contact with another. Commutator ripple, F_c , may be computed by the equation

$$F_c = \frac{\text{number of segments} \times \text{r.p.m.}}{60} \quad (181)$$

(2) *Slot ripple* is caused by the interruption of the magnetic field every time a slot in the armature passes a pole piece which results in a slight surge in the field changing the voltage induced in the coils. Slot ripple F_s may be computed by the equation

$$F_s = \frac{\text{Number of slots} \times \text{r.p.m.}}{60} \quad (182)$$

(3) *Brush sparking* is caused by an uneven contact between the commutator and the face of the brush. It cannot be predetermined by computation and can only be reduced in a given generator by proper operation of the machine and a smooth clean commutator and brushes.

Ripple filters for direct-current generators consist of inductance (L) in series with the line, and capacity (C) in shunt, or in parallel, to the line. Various combinations of ripple filters are shown in Fig. 402. All ripple filters are low-pass filters, that is, they attenuate all frequencies above their cut-off point and pass all frequencies below this point. Their effectiveness depends, therefore, on designing them so that their resonant frequency is as low as possible. This means that the L and C in the

¹ Condensed from "Filter Facts," a booklet from the Electric Specialty Company.

circuit must be as high in value as practical and economic conditions permit.

In circuit Fig. 402(a) the ripple voltage across the condenser is the ripple voltage across the tube which is the high-impedance load to which the condenser is connected. This may rise to several times the value of the initial voltage. The peak of this rise is reached at a frequency slightly less than resonance; beyond this point it decreases rapidly, soon becoming but a small fraction of the impressed voltage. This circuit, therefore, magnifies frequencies below its cut-off point (resonance) and greatly decreases those above this point. To be effective it must be designed with as low a cut-off point as possible which means the L and C values must be high. In the following paragraphs of this section the circuits referred to are in Fig. 402.

In circuit (b) we have circuit (a) with an inductance added in series with the load. This is called a T filter and is not as effective as circuit (a) because it increases the cut-off frequency and allows more ripple voltage to pass.

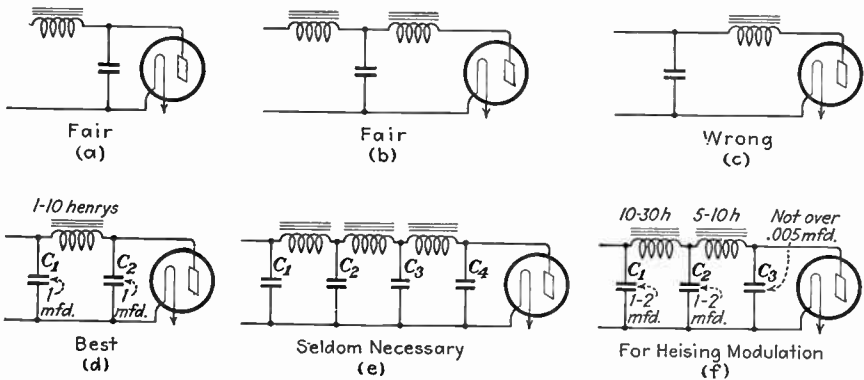


FIG. 402.—Generator ripple-filter circuits.

Circuit (c) is not a good low-pass filter for generators. The voltage across the condenser will be the full ripple voltage, and the ripple voltage across the plate will vary only as the impedance of the coil in series with it varies.

The π type of filter shown in diagram (d) is one of the simplest and most economical of filters. For the generator it is ideal. Properly built it is most effective of the "smoother" types. That is, its filtering effect is not critical. It functions at all frequencies above the cut-off or resonance point. The lower this cut-off point, and the sharper and more rapid the reduction beyond this point, the better is the filter. This means that condensers and chokes as large as is practical from an economic standpoint should be used.

A general idea of the functioning of this type may be obtained by considering it as divided into two parts, A and B . Part A is the condenser C_1 across the generator. Its effect on the plate circuit is small. Its effect on the minute ripples in the generator is tremendous. Part B , the choke and C_2 , takes what little disturbance is left and reduces it further as explained for circuit (a). One or two 1-mf. condensers and a 1- to 10-henry choke make a good filter for telegraphy and telephony work. Larger condensers and chokes will, of course, increase the filtering effect.

The single-section π filter just explained should be ample for most purposes. There are some combinations such as the belt-driven generator with its surges caused by belt slap, the separately excited generator with a large ripple in the exciting current, or a generator boosted with an alternating-current supply in series with it that

will need a little more elaborate filter. This may be accomplished by using a multi-section filter as shown in diagram (e). The first section reduces the disturbance a certain percentage. The next section reduces the ripple that the first section passes approximately the same percentage that the first section reduced the original ripple. The action continues with each section. The sections should be added, value for value. That is, the inside condensers will be twice the value of the outside ones, and the inductances should all be equal:

$$L_1 = L_2 = L_3; \quad C_1 = C_4; \quad C_2 = 2C_1; \quad C_3 = C_2.$$

If a π filter is used with Heising modulation, the filter should have added to it a small choke, 5 to 10 henrys, and a small condenser, not over 0.005 mf., to prevent the modulation frequencies from being short-circuited by the large filter condensers C_1 and C_2 . This circuit is shown in diagram (f).

Ripple filters, in addition to blocking out the ripple, also serve as a protective system guarding the generator against high-frequency kick-backs. Radio-frequency choke coils in the vacuum-tube plate circuits serve the same purpose by keeping radio-frequency currents out of the power supply.

V. ELECTRICAL INDICATING INSTRUMENTS

The radio engineer, technician, or operator spends so much of his time watching meters that a knowledge of their working principles is in order. The power-switchboard meters most commonly employed in radio stations may be grouped as follows: ammeters, voltmeters, frequency indicators, wattmeters, and ampere-hour meters. These instruments are designed to operate on direct or alternating currents.

378. Voltmeter for Continuous E.M.F.—Most direct-current voltmeters operate on the principle of the D'Arsonval movement which depends for its operation on the torque produced on a conductor by a field of force. In modern applications of this principle to voltmeters a rectangular coil of fine copper wire is fixed to revolve between the soft-iron core and pole pieces of a permanent magnet as shown in Fig. 403. This class of instrument commonly known as the *moving-coil* type is comparable in operating principle to a small electric motor.

The terminals of the moving coil are connected across the potential to be measured. The effect of the current flowing in the coil sets up a north and south pole in the coil which then tends to orient itself with respect to the magnetic poles of the permanent magnet between which it is fixed to revolve. The amount of twisting depends upon the current

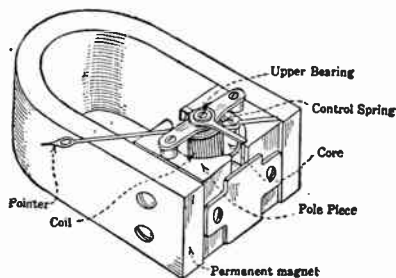


FIG. 403.—D'Arsonval movement for electrical indicating instrument.

flowing in the coil, and it is, therefore, possible to use it as a measuring instrument.

Because the voltmeter is *always* connected directly across, or in shunt to, the line, a *current-limiting resistor* must be connected in series with the moving coil. Generally for voltages up to 150 volts this resistance is mounted inside the meter. This resistance is of such a value that the amount of current flowing through the coil is negligible, and the movement of the needle, attached to the coil (see illustration) across the indicating scale, is uniform regardless of the potential value applied to the meter terminals. It is, therefore, possible to calibrate the meter to read any value of voltage up to its maximum capacity by merely changing the value of the resistance in series with the moving coil. Multi-range meters are in fact arranged with this series resistor tapped at different values to give the desired readings. The range values are so chosen that they are multiples or submultiples of one another and can be readily marked on the scale.

For voltages greater than 150 volts the series resistor is usually so large that it cannot be placed in the meter case and is generally supplied as a separate unit. For this type of meter, when of the portable type, it is the usual practice to supply several different values of resistances so that the range of the meter can be made flexible to suit conditions. These series resistors are called *multipliers* and a meter must be used only with the multiplier or multipliers for which it is calibrated, otherwise the readings will be erroneous.

A study of Fig. 403 will show how the meter is constructed. The pointer is attached to the moving coil as shown, and a small spring is provided which serves to bring the coil back so that, with no e.m.f. applied, the pointer points to zero on the scale. If a meter must be repaired for any reason, it generally requires recalibration at the same time.

As the D'Arsonval movement will give a forward deflection only when the current is flowing in a certain direction, the terminals of voltmeters employing this type of movement are marked with polarity signs. For this reason voltmeters may be used to determine the polarity of direct-current circuits.

Figuring Multiplier Values.—Suppose a certain D'Arsonval movement has a resistance of 2.5 ohms, and that it takes 0.030 amp. for full-scale deflection. Then, $E = I \times R = 2.5 \times 0.030 = 0.075$ volt, or 75 mv., for full-scale deflection of the movement, that is, without any resistance in series with the moving coil.

For example, suppose it is desired to utilize this movement as a voltmeter to read 150 volts, what must be the value of the multiplier?

The resistance value of the multiplier must be such that it will have a drop of $150 - 0.075$ or 149.925 volts across it when 0.030 amp. is flowing through it. Hence,

$$R = \frac{149.925}{0.030} = 49,975^{\omega}$$

It is useful for the radioman to know that any direct-current milliammeter can be converted into a voltmeter by connecting a suitable multiplier (R) in series with it. The value of the resistance may be found, as explained above. For example, suppose it is desired to convert a 0 to 5 milliammeter into a voltmeter to read 500 volts. Then, $R = E/I = 500/0.005 = 100,000$ -ohm multiplier necessary. When the range desired exceeds ten volts and the resistance of the meter is extremely low, compared to the total R , as is usually true, this small meter resistance can be neglected as is done here. For very accurate readings the converted meter would need to be calibrated against a standard meter.

The resistance of a voltmeter includes the resistance of the moving coil, plus total multiplier resistance, plus the resistance effect of any shunt resistances. Thus a 30,000-ohm voltmeter is a voltmeter the R of which measures 30,000 ohms between its terminal posts.

High-resistance voltmeters are used to measure voltages on circuits where the current drain is small. The reason for this is apparent when it is realized that any appreciable current drain by the meter makes a consequent change in the voltage of the circuit, and the meter would not show true operating voltages. Because of this, voltmeters for plate-supply and bias-voltage measurements usually have a resistance of about 1,000 ohms per volt, that is, a 0 to 50 voltmeter would have a resistance of 50,000 ohms. The ordinary portable voltmeter has a resistance which usually varies between 50 and 150 ohms per volt depending on the type of meter.

379. Ammeter for Direct Current. *a. Description.*—This instrument is constructed along the same lines as the voltmeter for direct current previously described, using the D'Arsonval movement, except that the moving coil of the ammeter is wound with a heavier wire and is of a lower resistance, usually from $\frac{1}{2}$ to 1 ohm in American instruments. The ammeter, as far as movement alone is considered, is a millivoltmeter, and, in fact, reads millivolts, which are directly proportional to the current flow. Ammeters can, therefore, be calibrated in amperes or milliamperes as desired.

The ammeter is always used in conjunction with a *shunt* which is a predetermined length of conductor of a specified resistance across which the ammeter is connected. The way in which the shunt and the ammeter are used in the circuit is shown in Fig. 404. The purpose of the shunt is to by-pass, or sidetrack around the meter, the greater part of the current being measured, thus making it unnecessary for the meter to carry much current in order to give an accurate reading.

The shunt is constructed by silver-soldering strips of a special alloy called *manganin* between two heavy copper lugs. The manganin strips act as the resistance across which a voltage drop occurs which voltage drop is measured by the millivoltmeter calibrated as an ammeter. Manganin strips are used, as it has been found that the resistance of this metal changes less than other metals when heated. The strip construction further increases heat radiation and reduces temperature effects on the resistance value of the shunt. It must be remembered that the voltage drop across the shunt is governed by the current flowing in the circuit and the resistance of the shunt. Therefore, if the resistance of

the shunt varied with temperature, a constant current would show up as a fluctuating one on the meter. The resistance of a manganin shunt does not vary with heat when properly designed for the load to be carried.

In most ammeters of the portable type the shunt is mounted within the instrument case in meters not reading over 25 amp. When the range is greater than this, a separate shunt to which the instrument is connected by leads is provided. Separate shunts are also provided for high-grade laboratory instruments. It is becoming more and more the favorite practice to use separate shunts in connection with switchboard-type ammeters. When this is done the main currents do not have to be taken off the most convenient and best routes, it being only necessary to put a suitable shunt in series with the line at any point and run leads to

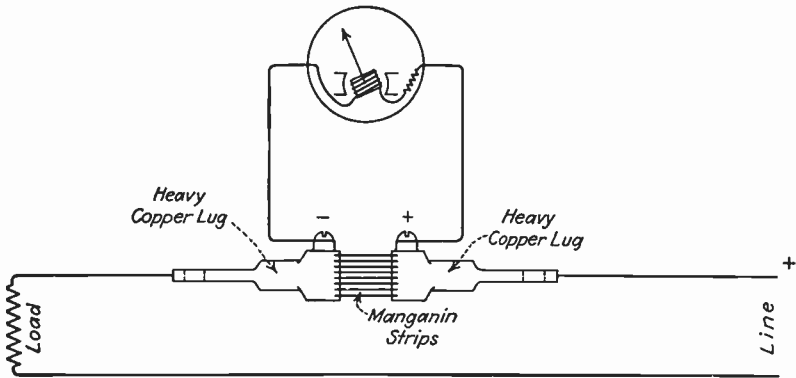


FIG. 404.—Shunt connected to ammeter.

it from an ammeter movement located in any convenient position on a switchboard or elsewhere. This practice also makes it possible to use one meter for various measurements, connecting it by a suitable rotary switch to the various shunts as required.

b. Figuring Shunt Values.—There are two factors which must be taken into consideration in the building of a shunt from the electrical-design standpoint. They are the resistance and the current-carrying capacity of the shunt. The resistance of the shunt is found by the equation $R = E/I$, where E equals the voltage necessary for a full-scale deflection, and I is the full-scale deflection in amperes desired on the meter when used in conjunction with the particular shunt to which the R value in ohms applies.

For example, if the voltage necessary to give full-scale deflection, as found by previous measurements, is 0.075 volt, and it is desired to have the ammeter read 30 amp., then the value of the shunt is $R = 0.075/30 = 0.0025$ ohm. If the resistance value of the shunt as thus found is inconvenient from a manufacturing standpoint, it is only necessary to change the series resistance used with the movement which would vary the E factor in the above equation. Hence by considering both the shunt value in ohms and the value of the series resistor a practical compromise may be found.

The current-carrying capacity of a shunt is found by the formula I^2R .

As the resistance of the meter must include the resistance of its leads, it is apparent that the length of these connecting leads cannot be materially changed after calibration of the meter without affecting its accuracy. The length of these leads is not critical, however, in an ammeter or milliammeter, and, under ordinary conditions, the accuracy of the meter is not affected by reasonable changes, provided extreme accuracy is not required.

c. A *current-limiting resistor* is used in series with the winding of the moving coil in nearly all ammeters. The value of this resistance varies with the type of meter. For switchboard purposes it is necessary to make the shunt loss low by developing only a small e.m.f. across it. It is necessary, therefore, to make the switchboard-ammeter movement sensitive to small potentials, and only a low resistance is used in series with the moving coil in this type of instrument. On the other hand, it is the practice to use a higher resistance in series with the moving coil of portable ammeters to reduce the temperature error in the movement, which is more important in portable service than reducing shunt loss.

d. The *direct-current milliammeter* is the typical ammeter movement described above with a higher resistance in its shunt and less in its series resistor.

The scale spacing on all direct-current ammeters and voltmeters is uniform from zero to maximum.

It is conventional to connect ammeters and milliammeters in the positive side of the line as shown in Fig. 404. Notice the relation between the positive and negative markings on the meter and on the line. It is seen that while the positive side of the meter is connected to the positive side of the line direct, the negative side of the meter is connected to the negative side of the line through the load.

It is sometimes desirable to make a shunt for an ammeter or milliammeter already on hand. The problem then is to figure out the resistance of the shunt to be used. The ratio is expressed as follows:

$$\text{Resistance of shunt} = \frac{\text{proportion of } I \text{ through meter}}{\text{proportion of } I \text{ through shunt}} \times R \text{ of meter} \quad (183)$$

For example, suppose it is desired to make a 0-1 milliammeter read to 10 ma., the resistance of the meter being assumed to be 30 ohms.

Then,

$$R \text{ of shunt} = \frac{1}{9} \times 30 = 3.33 \text{ ohms}$$

The resistance of the meter in the above equation means the resistance measured across the terminal posts of the meter. See also Sec. 11.

380. Voltmeter for Alternating E.M.F.—These instruments indicate the r.m.s. value of the alternating e.m.f. For radio-power purposes two types are most commonly used, namely, the *dynamometer type* and the *soft-iron-vane type*.

The dynamometer type depends for its action on the reaction produced between a movable coil and a stationary coil, or between a movable coil mounted between two fixed coils. This movement is called the *dynamometer*, and the general scheme is illustrated in Fig. 405(a).

The soft-iron-vane-type meter depends for its action on the deflection which takes place between two pieces of soft iron bent concentrically

and placed within a coil. The scheme is shown in Fig. 406. When the coil is energized, it magnetizes both pieces of iron in the same direction with the result that they repel one another resulting in a movement of the movable coil. The extent of the movement depends on the amount of

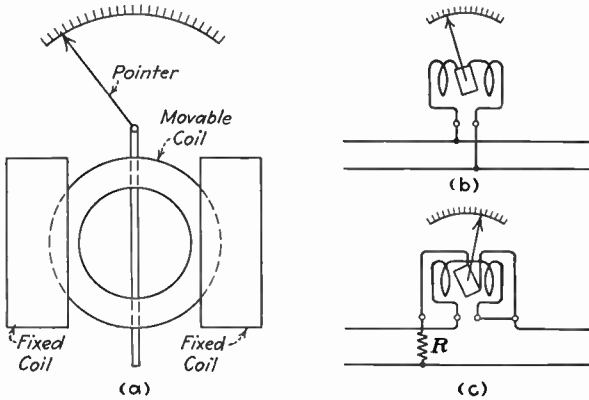


Fig. 405.—(a) Dynamometer movement; (b) series-connection method for small currents; (c) wattmeter connection for large currents.

excitation produced by the coil. A pointer is attached to the movable element which passes in front of a calibrated scale. As the reversals of magnetization due to the alternating current occur in the coil, the iron vanes are also reversed in polarity and the deflection factor remains constant regardless of polarity. It is possible, therefore, to use this type of movement on an alternating-current circuit.

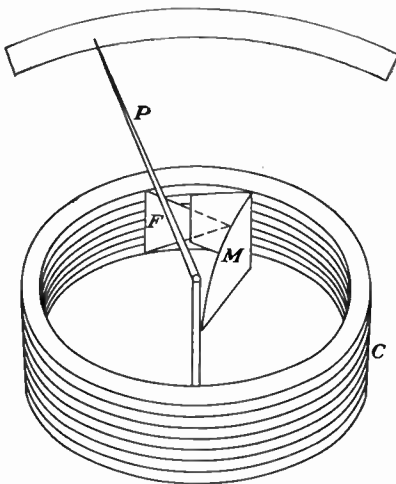


Fig. 406.—Iron-vane-type meter.

non-uniform. For example, on an alternating-current instrument of the types just described the spacing between the scale units 3 and 4 is approximately one quarter the spacing between the units 15 and 16, assuming, of course, the instrument reads that high.

Both the dynamometer-type and the soft-iron-vane-type instruments are limited to use on commercial frequencies only and cannot be used to measure radio frequencies owing to the errors introduced when they are used on high frequencies.

The deflection of alternating-current instruments is non-uniform and tends to be proportional to the square of the current flowing through the meter. The indicating scale on these instruments is therefore of necessity

Another type of voltmeter called the *Thompson inclined-coil type* is very similar to the dynamometer type except that the stationary and movable coils are inclined, so that the plane of the fixed coils makes an angle of about 45 degrees with the shaft of the movable coil. This inclination of the coils tends to make the scale reading more uniform.

381. Ammeters for alternating-current circuits divide themselves into several types and classes ranging from the dynamometer and iron-vane types for commercial frequencies (usually not over 600 cycles) to the hot-wire and thermocouple types used for alternating currents of any frequency.

Ammeters utilizing the dynamometer movement (see Sec. 380) have the coils connected in series [Fig. 405(b)] when they are intended for current values up to 1 amp.; for values higher than 1 amp. the coils are connected on the wattmeter principle where the movable coil is used to register the potential effect and is connected across the line in series with a current-limiting resistor, and the stationary windings are used to utilize the current effect and are connected in series, the pair being connected in series with the line. This method of connection is shown in Fig. 405(c) and is called the *wattmeter connection*.

Another type of ammeter is the Thompson inclined-coil type which is classed as an iron-vane-type meter but differs somewhat in construction from the Weston type of iron-vane meter (Fig. 406). The Thompson inclined-coil ammeter is similar to the Thompson inclined-coil voltmeter for alternating-current circuits, except that the moving coil has been replaced by an iron vane mounted at an angle to the shaft. When a current passes through the coil the vane tends to arrange itself parallel to the field set up by the coil.

382. Thermal Ammeters. *a. General Considerations.*—The thermocouple ammeter and the hot-wire ammeter are the two types of thermal meters used in radio. Certain considerations and characteristics belong to these meters as a class. It was soon found by early radio engineers that the ordinary moving-coil or iron-vane type of meter could not be used to measure radio-frequency currents accurately. Even if these meters were accurate at high frequencies, the inductance of their windings would change the oscillatory period or frequency of the radio circuit.

The hot-wire ammeter was then introduced for this work and permitted current measurements because it operated on any frequency, and because it had no winding it introduced no change into the radio circuit, the resistance of the platinum-silver wire being negligible.

It was then found that the accuracy of the moving-coil type of direct-current meter in conjunction with the e.m.f.-producing characteristic of the thermocouple unit could be used to measure radio-frequency currents. This resulted in the thermocouple ammeter which is now the

type used almost entirely for radio-frequency-current measurement purposes.

In addition to its great accuracy and ruggedness, the thermocouple meter introduces no appreciable added inductance, capacity, or resistance into the circuit under measurement. It is for these reasons, in addition to its ability to measure all frequencies, that the thermocouple ammeter is used in radio-frequency circuits.

All thermal meters may be used on either direct or alternating current and they are independent of frequency.

b. *Thermocouple Ammeters.*—Certain metals, such as steel and constantan, bismuth and antimony, will produce a constant e.m.f. if brought under contact pressure at high temperatures. Such a combination of metals for this purpose is called a *thermocouple unit*. The heat

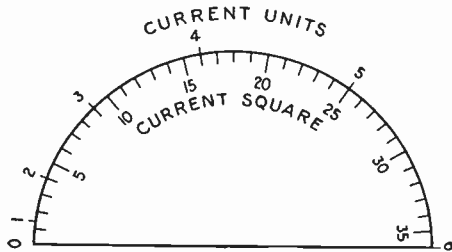


FIG. 407.—Meter scale.

which produces the e.m.f. is in turn produced at a rate which is proportional to the square of the current flowing through the thermocouple.

A direct-current millivoltmeter may be connected across the thermocouple to measure the e.m.f. produced in it. As the e.m.f. is proportional to the heating of the thermocouple, and as the heating is increased as the square of the current flowing through the thermocouple, then the e.m.f. must be increased as the square of the current flowing through the thermocouple. The millivoltmeter may, therefore, be calibrated in evenly spaced heat units across its scale, or it may be calibrated in amperes or milliamperes, or a combination scale may be used as shown in Fig. 407. If the scale is calibrated in evenly spaced units, the meter is called a *current-square meter*; if calibrated in current units it is simply an *ammeter* or *milliammeter*.

Current-square meters are valuable when small variations in current are to be noted, and this type of meter is useful on wavemeters and in other places where resonance and amount of deflection are more desired to be known than actual current values. It is evident that the actual current is the square root of the value in units indicated on the scale of the current-square meter.

The thermocouple ammeter will read either direct or alternating current of any frequency. Its chief use, however, is for measurements

in radio-frequency circuits. As the thermocouple and the indicating meter are two separate and distinct devices, they may be located some distance apart and connected electrically if desired. Generally, the thermocouple unit is placed within the meter case unless the heat generated is so great as to make an outside location desirable, or the point where the measurement is to be made is remote from a convenient position for the meter.

The construction of the thermocouple unit is illustrated schematically in Fig. 408(a) where *hh* is the heater wire and *tc* are the two metals required for the thermocouple. The entire arrangement with terminals for the meter and the circuit forms the thermocouple unit. Particular

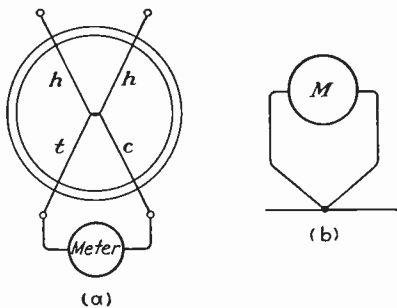


FIG. 408.—(a) Thermocouple and meter; (b) correct symbol.

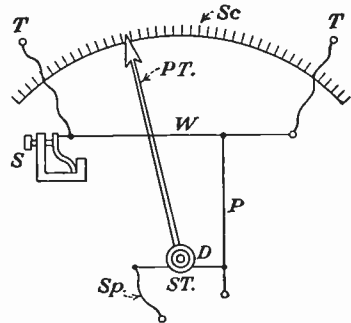


FIG. 409.—Hot-wire-ammeter mechanism.

attention is called to the correct method of drawing the symbol for the thermocouple ammeter which is properly shown in Fig. 408(b).

c. Hot-wire Ammeter.—Another type of thermal meter is the hot-wire ammeter which utilizes the expansion characteristic of a thin platinum-silver-alloy wire, when subjected to the heating effect of a current, to indicate current flow. The principle of the mechanism of this type of meter is shown in Fig. 409. The hot-wire ammeter is not very rugged owing to the fragility of the platinum-silver wire. In addition to this it is not very accurate and may lose its calibration. Also, owing to the thermal-expansion characteristic of the platinum-silver wire, a number of seconds must elapse after the application of the current before the meter will register. For these reasons and others, the hot-wire ammeter has been largely replaced by the thermocouple type of meter. The hot-wire ammeter does, however, still find some use in wavemeters for determining the resonance point and also for measuring the radiation, filament, and plate current in experimental transmitting stations where economy prohibits the use of a thermocouple meter.

The operation of the hot-wire instrument may be explained as follows: When the current passes through the platinum-silver wire, *W*, it heats the wire, causing it to slacken; then the pointer, *PT*, being mechanically con-

nected to the wire by a silk thread $P-ST$ wound around a drum, D , on the pointer axle, is propelled over the scale Sc in accordance with the amount of slackening of the hot wire. The scale may be either a current-square scale, or it may be calibrated directly in amperes. A screw S is provided to adjust the pointer to zero on the scale, and, as the wire is affected by room temperature, this is often necessary. A series resistance (multiplier) may be used to extend the range of a hot-wire ammeter and is connected in series with the meter and the line being measured. T and T' are terminals.

383. Wattmeters.—There are two types of wattmeters, the dynamometer type and the induction type. As the dynamometer type is most commonly used, this type will be described here. The wattmeter indicates the *rate* of energy expenditure or *power* being consumed in a circuit at any instant. The wattmeter should not be confused with the watt-hour meter which registers the total *amount* of energy consumed in a circuit. The ampere-hour meter, described in Sec. 418, is a type of watt-hour meter.

The dynamometer-type movement used in the wattmeter is the same as that used in the dynamometer-type voltmeter, described previously and which is illustrated in Fig. 405(a). The fixed and movable coils are connected in accordance with the wattmeter connection shown in (c) of the same figure.

If an alternating current circuit contains resistance only, it has unity power factor, and the power expended in watts is found by the simple equation $W = E \times I$, but, if the circuit contains inductance or capacity (or the equivalent), the power factor must be taken into consideration. The equation then becomes $W = E \times I \times \cos \theta$. The wattmeter automatically takes these three factors into account, and the value indicated by the pointer on the meter scale is a true power reading.

In representing the wattmeter (or a dynamometer-type meter in any other use) the current-limiting resistor [R , Fig. 405(c)] in series with the shunt circuit should always be shown.

384. Frequency Indicators.—The preferred term *frequency indicator* rather than *frequency meter* is used here to differentiate between the instrument used on power lines to measure the frequency of alternating c.m.f.s. at power frequencies and the conventional wavemeter circuit commonly known as a *frequency meter*, and used to measure the frequency of radio-frequency currents.

The frequency indicator provides a scale and a movable pointer from which may be read directly the frequency of the alternating-current circuit to which it is connected. There are other types, such as the vibrating-reed type, which are built on a different principle but which need not be considered here. The frequency indicator met with in everyday practice looks like a voltmeter, ammeter, or any other conventional electrical indicating instrument.

A common type of frequency indicator is the Westinghouse induction type, the general scheme of which is shown in Fig. 410. Its construction and operation are explained as follows: Two coils *A* and *B* act on the rotatable disk *G* tending to turn it in opposite directions when a current flows through the coils. A non-inductive resistance *R* is connected in series with the *A* coil, and an inductive reactance X_L is connected in series with the other coil, *B*. The amount of current flowing in the coil *B* is controlled by the impedance of the reactance X_L which varies with frequency. Hence, the effect of the two coils on the disk is unbalanced and the disk tends to orient itself to a definite position in accordance with the frequency of the applied alternating current. *PP* are the poles of the field magnet.

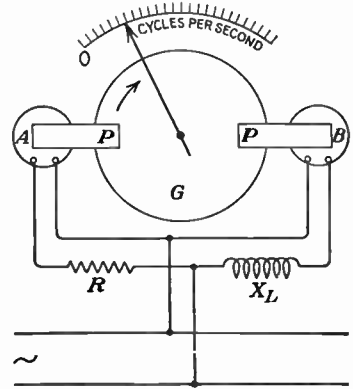


FIG. 410.—Induction-type frequency indicator (Westinghouse).

Note: The ampere-hour meter is described in Sec. 418, next chapter.

385. The Ohmmeter.—The simplest type of direct-reading ohmmeter and the one generally used for radio-circuit testing and resistance measur-

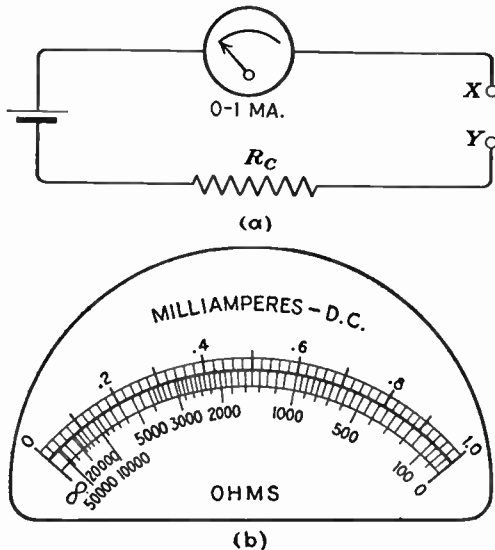


FIG. 411.—(a) Ohmmeter circuit; (b) ohmmeter scale.

ing consists of a direct-current milliammeter, a known resistance (R_c), and a dry cell or cells for supplying e.m.f., arranged in a circuit as shown in Fig. 411(a). The unknown resistance, or the circuit, the continuity of

which is to be tested, is connected across the terminals XY . With XY short-circuited, the resistor R_c is of such value that the applied e.m.f. causes a full-scale deflection on the meter. It is evident then that any resistance introduced across XY will cause the meter to read less than full scale, the amount less depending on the amount of resistance introduced at XY . The ohmmeter, therefore, reads backward, that is, zero ohm is at the right-hand end of the scale.

A typical ohmmeter scale is shown in Fig. 411(b). Using this scale in conjunction with a Weston type 301, 0 to 1 milliammeter, resistances up to 50,000 ohms may be read directly on the scale when the applied e.m.f. is 1.5 volts and the value of R_c is 1,500 ohms. The range of the meter may be extended by increasing the value of R_c . The factor by which R_c is multiplied then becomes the multiplying factor for the scale. For example, if R_c is increased 15 times ($1,500 \times 15$) to 22,500 ohms, then the scale indication must be multiplied by 15, giving a maximum range of ($15 \times 50,000$) 750,000 ohms. The battery voltage must always be great enough to operate the meter. Hence, when very high resistances are to be measured, the battery e.m.f. must be increased proportionately. For certain radio-circuit tests an ohmmeter is indispensable.

Another simple and very accurate method for determining unknown resistance values is the voltmeter method. If the resistance of the voltmeter is known,¹ the value of any unknown resistance may be determined by the following procedure:

Let it be assumed that the voltmeter has a resistance of 100,000 ohms and a full-scale reading of 150 volts. That is, if the voltmeter is connected across an e.m.f. of 150 volts, the meter will give a full-scale deflection of 150 volts. Now, if an unknown resistance is connected in series with the instrument under these conditions, the reading on the voltmeter scale will obviously indicate less than 150 volts, owing to the IR drop across the unknown resistance. If, therefore, the insertion of the unknown resistance drops the scale reading to, say, 125 volts, then the value of the unknown resistance may be readily computed from the expression,

$$R_x = R_m \times \left(\frac{E_1}{E_2} - 1 \right) \text{ ohms}$$

or in this instance,

$$R_x = 100,000 \times \left(\frac{150}{125} - 1 \right) = 20,000^\omega$$

This method is extremely simple and accurate, particularly where high values of unknown resistances are to be determined.

¹ The resistance of any voltmeter in question may be obtained from the manufacturer or by measurement with a standard ohmmeter.

CHAPTER XV

STORAGE BATTERIES¹

Storage batteries are used extensively in all classes of radio stations. They are made for both portable and stationary use, and, as far as the lead-acid type is concerned, the use for which a battery is designed determines the specific gravity of its electrolyte. Generally speaking all lead-acid-type batteries for stationary use are of the 1.210 specific-gravity type; whereas those designed for portable use have a gravity of 1.275. The higher gravity in the portable battery may be explained by the restricted area of the plates, compared with the stationary type, and the reduction in quantity of electrolyte, which makes a stronger solution necessary. The stationary type of storage battery is used at some broadcasting stations and at high-powered radio-communication stations located ashore where an emergency power supply is desired. The portable type of battery is used on mobile stations where space and operating conditions make the installation of a stationary type impractical.

386. General Battery Considerations.—Storage batteries are sometimes used in connection with broadcast-studio equipment where the batteries are floated across a generator in which position they act as a filter condenser to absorb any ripple and to provide a source of emergency power for use should the normal power supply fail. In this way the battery is always in the circuit, and in the event of a generator failure the battery is across the line and no interruption is noticed. However, its use as a ripple filter in broadcast stations is perhaps more important in practical operation than is its ability to furnish emergency power, as power failures are comparatively rare occurrences.

For marine use a 60-cell storage battery of the portable type is generally supplied. This battery provides current to operate the main radio motor generator in case of failure of the ship's generator. An emergency lighting circuit is also usually connected to this battery. Since the development of low-power vacuum-tube transmitters capable of fulfilling the requirements of the law as to transmission range, companies manufacturing radio equipment have found that a motor generator driven by a 12-volt motor is satisfactory. A transmitter of this type is described in Sec. 253. Under certain conditions, therefore, the emergency radio battery need only be capable of delivering 12 volts.

¹ Much of the material in this chapter pertaining to the practical operation of the lead storage battery has been taken, by permission, from instruction booklets issued by the Electric Storage Battery Company.

There are two distinct types of storage batteries, the lead-acid type and the nickel-iron-alkaline type, the latter being known as the *Edison battery*. Whenever reference is made to a storage battery, or cell, without specifying to which type it belongs, the lead-acid type is meant.

The use of the nickel-iron-alkaline battery in radio has been principally confined to installations aboard ship for use as an emergency battery to comply with the law which requires a source of emergency power for operation of the ship's radio transmitter.

I. THE LEAD-ACID BATTERY

387. The Lead-acid Cell.—Storage batteries produce an e.m.f. as a result of chemical action after they have been subjected to an initial charge. This action takes place between the active material of the positive and negative plates and the electrolyte, which in the lead cell is a dilute solution of sulphuric acid. The active material of the positive plates is lead dioxide¹ (PbO_2) which is always chocolate brown in color when in its normal condition. The active material of the negative plates is commercially pure sponge lead (Pb) which is pearl gray in color when the plate is in a normal condition. If the negative plate is white, slate gray, or spotted, an abnormal condition is indicated.

The great majority of lead cells used in radio-station service in the United States are manufactured by the Electric Storage Battery Company (Exide) of Philadelphia. This company manufactures a number of general types of cells each differing in the types of positive and negative plates used. Two of these types are in common use in radio stations; they are (a) the chloride accumulator² and (b) the Exide battery. The first named is a stationary type (Fig. 415), while the latter is a portable battery [Fig. 414(b)].

388. Positive Plates.—There are two types of positive plates, the Planté and the Faure, the former being used in stationary and the latter in portable batteries.

a. Planté Type.—This type of plate as manufactured in the United States by Exide is known as the *Manchester positive* plate and consists of a grid of cast lead-antimony alloy which resists both electrical and mechanical strain and, therefore, retains its strength, shape, and dimensions. This grid is provided with circular openings into which are forced, by hydraulic pressure, the rosettes or buttons of soft lead which constitute the active material of the plate. The construction of the Manchester plate is clearly seen in the cell shown in Fig. 412. After being forced into place in the grid, the buttons are subjected to the "forming" process, whereby the active material, lead dioxide, is developed electro-chemically

¹ Lead dioxide is sometimes erroneously called lead peroxide.

² Chloride accumulator is a misnomer, there being no chloride whatever used in the lead cell.

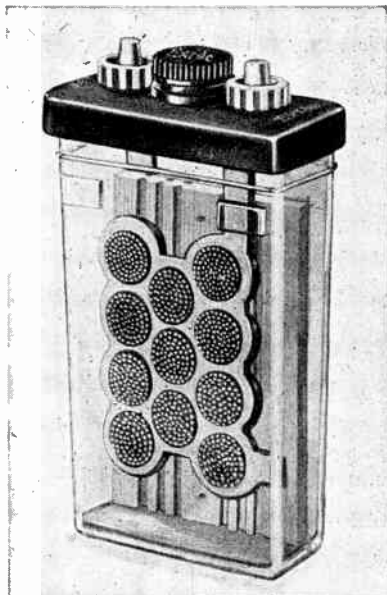


FIG. 412.—Two-plate lead cell showing Manchester positive plate. (Courtesy of Electric Storage Battery Company.)

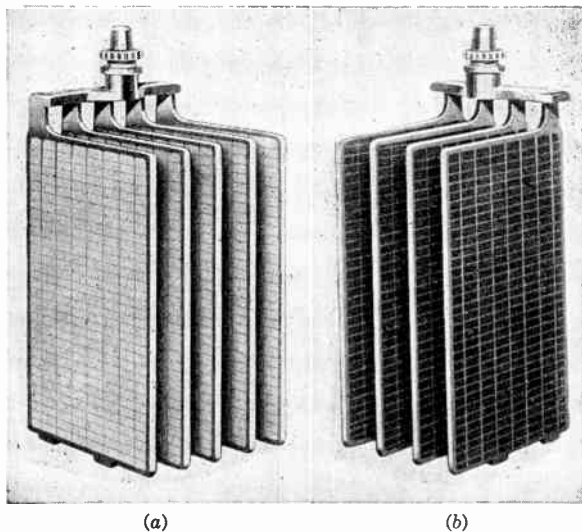


FIG. 413.—Constituent parts of lead cell element: (a) negative group; (b) positive group. (Courtesy of Electric Storage Battery Company.)

on the transverse surfaces. The fact that there is available a supply of solid lead in the rosettes or buttons that can be converted into active material as needed accounts for the long life for which this type of plate is noted.

b. Faure Type.—This is commonly known as the *pasted* or *flat* type of plate and consists of a lattice-like grid of stiff lead alloy which supports the active material. The grid is “pasted” with an oxide of lead made into a paste of special composition which sets in drying, like cement. The plates are then subjected to an electro-chemical process which converts the oxide of lead to dioxide of lead which is brown in color. This type of plate is shown in Fig. 413(b).

c. Exide-ironclad positive.—This is a modification of the Faure-type plate and consists of rows of hard-rubber tubes which are horizontally slotted and which contain the active material and permit free access of the electrolyte. This type of plate is visible in the cell shown in Fig. 414(a) and permits current drains, as high as 1,000 amp., without injury or overheating of the cell.

389. Negative Plates.—While negative plates may be constructed along the lines of the Planté type, it is the practice to construct all negative plates by the Faure method utilizing two methods of grid construction, the lattice or flat type and the box type.

The lattice-type negative plate consists of a lattice-like grid on which is pasted the active material of the negative plate, sponge lead. This plate is shown in Fig. 413(a), and is similar in construction to the Faure-type positive plate shown in Fig. 413(b).

The box-type negative plate consists of a lead-antimony-alloy grid which is formed of horizontal and vertical ribs, spaced about 1 in. apart, forming pockets which are closed on both sides with perforated sheet lead, in which pockets the active material is permanently held in place. This construction is clearly shown in Fig. 415. The active material consists of compressed blocks of lead oxide which are later formed into the active material of the negative plate. With this type of construction the grid is strong and rigid and is not affected by the electrolyte, while the active material is exceedingly porous and finely divided and retains this porosity during service.

390. The active material must be porous enough to allow a sufficient amount of acid to reach the active elements at all times regardless of the rate of discharge, except that this rate must always be within the rated limits of the cell.

The arrangement of the active material is dependent on whether the battery is to supply power at a high or low rate of discharge. If the rate of discharge is to be high, the material must be spread in a thin layer over a large area so as not to require a great depth of penetration by the acid in order to create the chemical action necessary to the produc-

tion of an e.m.f. If the rate of discharge is not relatively high a greater depth of active material, measured from the surface where the electrolyte makes first contact, may be applied as ample time is then allowed for acid action and penetration.

The quantity of active material required depends upon its arrangement; if a high-capacity cell is required, the quantity of active material must be increased proportionately, and vice versa.

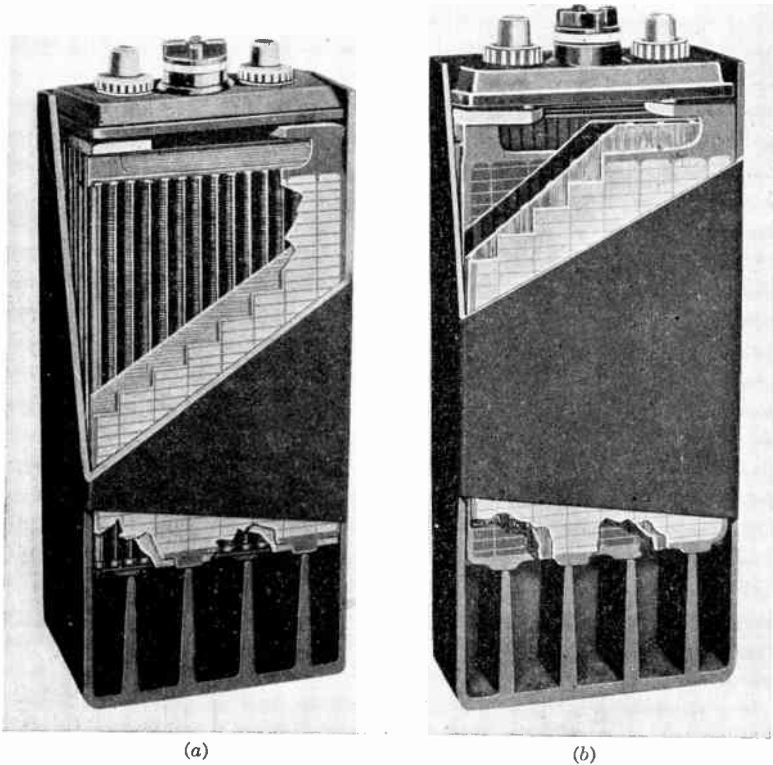


FIG. 414.—Types of lead batteries: (a) Ironclad Exide; (b) Exide portable. (Courtesy of Electric Storage Battery Company.)

391. The electrolyte used in lead-acid batteries is made by mixing sulphuric acid (H_2SO_4) and water (H_2O). The amount of each depends upon the gravity desired, which in turn depends upon whether the stationary or portable type of battery is used, the former utilizing a 1.210 specific-gravity electrolyte, and the latter, a 1.275 gravity electrolyte. The proportion of water to acid in an electrolyte with a specific gravity of 1.210 is 4:1 by volume, while the proportion for an electrolyte having a specific gravity of 1.275 is 3:1 by volume. This assumes the

use of commercially pure (93.5 per cent) sulphuric acid, having a specific gravity of 1.835.

All specific-gravity readings mentioned are at 80°F. when referring to 1.275 specific gravity, or 70 degrees when referring to 1.210 specific gravity. A plus or minus variation in the specific gravity of fully charged cells of 10 degrees in gravity is tolerated. In other words, when a cell having a specific gravity of 1.275 is spoken of, it usually means a reading ranging from 1.270 to 1.280.

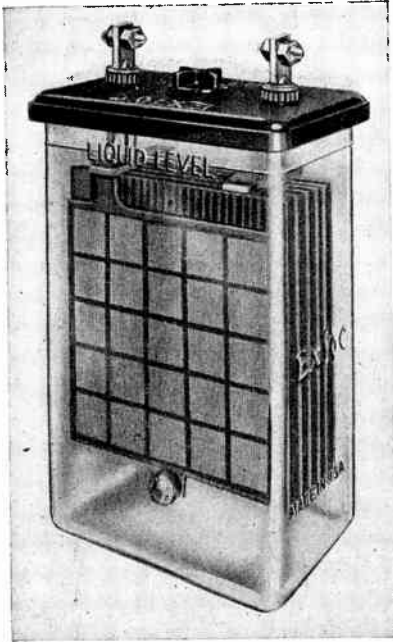


FIG. 415.—Stationary-type lead cell.
(Courtesy of Electric Storage Battery Company.)

The electrolyte should always be kept to the level specified for the battery, usually $\frac{1}{2}$ inch or more above the tops of the plates depending on the type of cell used. To maintain this level, never add anything but pure distilled water, or tap water which has been approved by the battery manufacturer as being satisfactory for this purpose. The reason for this is twofold: acid does not evaporate and, therefore, a lowering of the electrolyte level indicates evaporation of water; and impure water may cause local action in the cell with its consequent losses. Electrolyte may be lost by excessive spraying or gassing on account of too high a charging rate, and for this reason filling plugs must only be removed when taking a gravity reading or adding water.

The only time when acid is added to a cell is when some of the electrolyte is spilled out and cannot be recovered. This might happen during shipment, installation, or overhauling.

Every lead cell works best with an electrolyte of a certain density or specific gravity which is determined by exhaustive tests by the manufacturer. An electrolyte of a higher specific gravity will raise the voltage of the cell and consequently its capacity, but over a period of time the life of the cell will be shortened by such overloading. It is always advisable to work a cell with the electrolyte at the specific gravity recommended by the manufacturer.

As the specific gravity of an electrolyte varies with temperature, so also does the capacity of the cell vary with temperature. Generally speaking, the capacity of a cell varies directly with change in temperature, rising as the temperature rises, and lowering as the temperature lowers.

When batteries are used in very hot temperatures, better results are generally obtained by using an electrolyte of a lower specific gravity than would be required in cooler places.

Mixing Electrolyte.—Only commercially pure sulphuric acid should be used for storage-battery purposes. Do not confuse commercially pure acid with “full-strength” or “concentrated” acid; there may be a considerable difference which might work injury on the battery.

If electrolyte of the proper strength is not on hand it may be prepared from commercially pure sulphuric acid by mixing the acid with approved water. The acid may be any strength, provided it is stronger than the electrolyte desired. Special care must be taken in handling acid when it is stronger than 1.400 and extreme care when it is stronger than 1.600 to avoid personal injury.

The proportion of acid to water depends on the strength of acid used. When concentrated acid with a specific gravity of 1.835 is used, an electrolyte of 1.210 specific gravity contains approximately 4 parts water to 1 part acid by volume; whereas an electrolyte with a specific gravity of 1.275 requires not quite, but approximately, 3 parts water to 1 part acid by volume. When mixing take the following precautions:

1. Use a Pyrex glass, glass, china, earthenware, or lead vessel; never a metallic container other than lead. Rubber may be used, except when mixing acid of 1.600 gravity or stronger.

2. Slowly and carefully pour the acid into the water; not the water into the acid, as this might cause sputtering and, therefore, personal injury. Interrupt the pouring frequently, and stir thoroughly with a clean wooden paddle. Discard the paddle after using.

3. The gravity of the mixture will rise as the mixture cools. Therefore, if the mixing is completed with the temperature between 90°F. (32°C.) and 105°F. (40°C.), it should at this time read 0.005 to 0.010 specific gravity lower than desired; if temperature is between 105°F. (40°C.) and 120°F. (49°C.), it should read 0.010 to 0.015 specific gravity lower, etc.

4. Allow mixture to cool below 90°F. (32°C.) before adding it to the battery.

The temperature of the electrolyte or acid may be taken by a special thermometer supplied by the battery manufacturer which is shaped like a hydrometer and weighted at the bottom with lead granules. It is designed to float in the liquid being tested and is read like any ordinary thermometer.

392. Construction of Cells.—Cells are constructed in accordance with the cell type, the details of which are given in the table which follows.

Cell type	Positive plate	Negative plate	Type of service	Specific gravity
Chloride accumulator	Manchester-Planté type	Box type, Faure	Stationary land stations. Broadcasting not portable	1.210
Exide.....	Faure pasted-lattice type	Faure pasted-lattice type	Portable and stationary, usually mobile station use	1.275 for portable; 1.210 for stationary
Exide-ironclad....	Slotted tubes containing active material, Faure type	Faure pasted-lattice type	Lifeboat radio equipments	1.275

A study of the illustrations of the lead-acid batteries shown in this chapter will give the reader an excellent idea of how the cells are constructed. Referring to Fig. 413 it will be noted that each plate has a lug at the top. This illustration shows the Faure-type pasted plates, but cells utilizing Manchester plates have their plates arranged the same way. Regardless of the type of cell, a certain number of positive plates are "lead burned" to an alloy strap, forming a positive *group*. The negative group is assembled in the same way, except that there is always

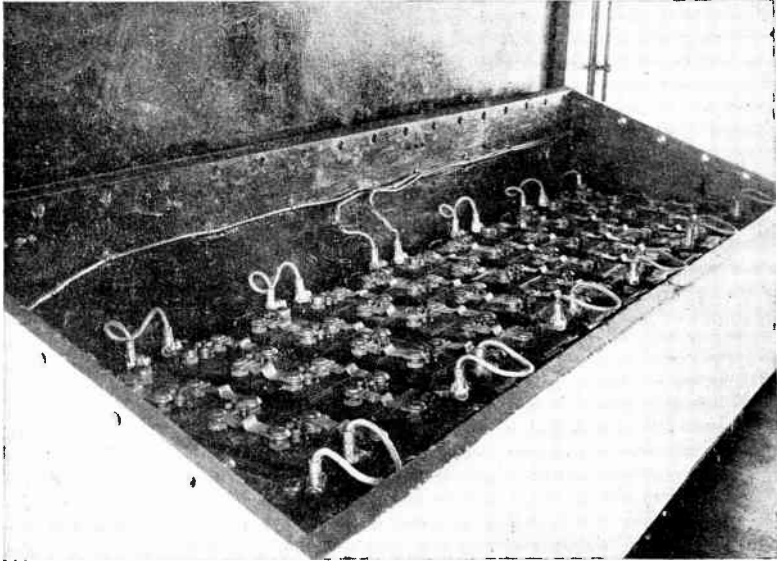


FIG. 416.—Lead-battery installation aboard ship. (Courtesy of Electric Storage Battery Company.)

one more negative plate than there are positive plates, so that when the plates are interlocked the outside plates are always negative. The exception to this rule is the two-plate glass-jar type shown in Fig. 412.

In order to insulate the positive group from the negative group there is inserted between the plates of most of the closed-jar types a grooved-wood separator, the grooved side of which goes against the positive plate and the smooth side of which goes against the negative plate. In addition to this wooden separator there is placed, in certain types of cells, a perforated or porous rubber sheet between the wooden separator and the positive plate. The entire assembly of plates and separators is known as an *element*, and the element is inserted in a glass or rubber *jar* which contains the electrolyte. Ample space is allowed in the bottom of the jar for the accumulation of sediment, and normally this space will hold all of the sediment deposited during the life of the cell. It is, therefore, unnecessary to disassemble a modern lead-acid cell to remove sediment.

The chloride-accumulator and Exide types of cells designed for stationary service are supplied in sealed glass jars. The Exide portable-type battery is supplied in a sealed rubber-jar container. In the sealed-jar types the necessary holes are provided in the tops for the terminals and for the pouring in or taking out of the electrolyte.

In the old-type open-glass-jar battery the plates are suspended from the sides of the jar which rests upon a tray filled with sand. This tray is employed to catch any creepage of electrolyte which may occur and so protect the wooden racks on which the cells are mounted. The sand acts as a cushion between the glass surfaces of jar and tray and thus prevents breakage which might otherwise occur. With the sealed jar, creepage is prevented by the sealed cover, which makes sand trays unnecessary. When the open-jar-type battery is employed, a separate battery room is necessary because of the destructive acid spray present.

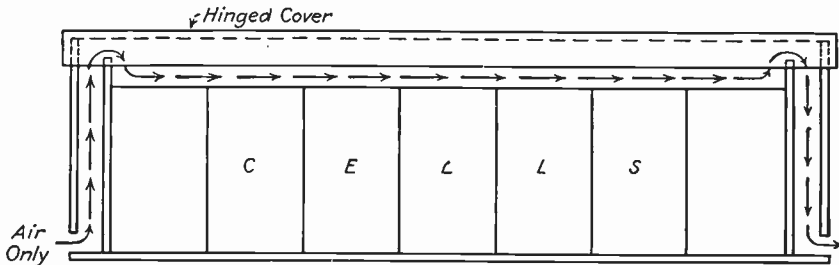


FIG. 417.—Plan of battery-housing box for shipboard use.

For stationary service the cells are mounted on wooden racks, in tiers if necessary. On shipboard, where the portable type of battery is used, the cells are mounted together in strong supporting boxes which are then called *trays of cells*. A typical shipboard installation of 60 cells is shown in Fig. 416. The large box which houses the installation is strongly built for marine use and is coated on the inside with an acid-resisting paint as are also the individual trays.

It is necessary that ventilation be provided in the battery-housing box. In the box shown in the illustration, holes are drilled around the top of the sides of the box to pass the air. This type of box should be opened during charging periods, if the weather permits, as good ventilation is then assured. A later type of box is built according to the plan shown in Fig. 417. Openings are provided in both ends of the box near the bottom and the air can take the route indicated by the arrows even if the cover is down. Sea water or spray could not reach the cells housed in this type of box unless the batteries were submerged.

The purpose of the battery housing is, of course, to protect the batteries from the weather and from salt water and spray. Salt water contains chlorine which is injurious to the batteries.

393. Chemical Action in a Lead Cell.—The chemical action in a lead cell during one cycle of charge and discharge is shown in Fig. 418. Four periods are shown in which the conditions existing in the cell from the time it is fully charged until it has been completely discharged and put on charge again are represented.

Period 1 shows the charged cell with the electrolyte at the proper specific gravity, at which time all of the acid is mixed with the water and the specific gravity reading is maximum. The negative and positive plates are in their natural condition having only their active material on them. The cell is as originally assembled, except that litharge (PbO) pasted on the plates has been formed into active materials, electrolyte is present, and the cell is fully charged.

Period 2 shows the cell as it starts to discharge. The acid is leaving the electrolyte and going into and combining with the active material of the plates. As it does this it reduces the active material of the positive and negative plates to lead sulphate,

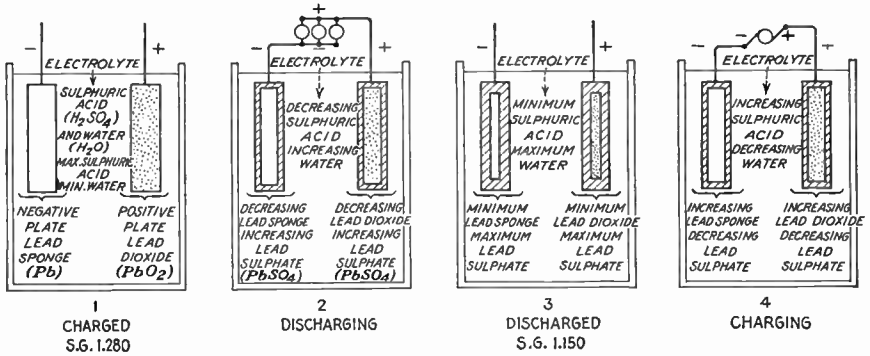


Fig. 418.—Chemical action in a lead cell.

decreasing the sponge lead and lead dioxide. This continues until nearly all of the acid has gone from the electrolyte into the active material of the plates. The specific gravity of the electrolyte becomes lower as the acid leaves the water and enters into chemical combination with the active materials. It is possible to measure the extent of this action by taking a reading of the specific gravity of the electrolyte and thus determine the state of charge of the lead cell. When so much of the acid has left the water that the specific gravity has dropped approximately 100 points from the charged reading, the cell is considered as being discharged.

Period 3 is the cell in a discharged condition when the electrolyte consists of a minimum amount of acid and a maximum amount of water. The porous active material is now clogged up and coated with lead sulphate and the action of the acid on the active materials of the plates is considerably slowed down which causes the voltage of the cell to drop. Although in this condition the plates are sulphated, their condition is not serious, that is, the sulphate can be removed; and it is easily removed if the battery is placed on charge shortly after discharge and before the sulphate hardens.

Period 4 shows the cell on charge. The chemical action taking place now is the reverse of that of period 2, the acid is coming out of its combination with the active material and is being restored to the electrolyte. The lead sulphate is changing back to its original form of sponge lead and lead dioxide. The cell is not fully charged until all, or nearly all, of the lead sulphate has gone from the plates. The life of a

lead cell is often shortened because the cell is given only a partial charge; that is, all of the sulphate is not removed from the plates and it becomes hard. This hardened sulphate tends to clog the active material, which is ordinarily porous, so that the acid cannot make sufficient penetration to cause the cell to operate properly. Therefore, the capacity of the cell is reduced by this hardened sulphate, and the cell is (*sulphated*.)

The combined reaction is expressed: $\text{PbO}_2 + \text{Pb} + 2\text{H}_2\text{SO}_4 \rightleftharpoons 2\text{PbSO}_4 + 2\text{H}_2\text{O}$.

394. Specific Gravity and the Hydrometer. *a. Specific gravity*, when applied to liquids, is a convenient method of comparing the weight of a given volume of liquid with a like volume of water. This gives a numerical value which means that the liquid, to which the value applies as a specific gravity, is that many times heavier than water.

$$\text{Specific gravity} = \frac{\text{weight of liquid}}{\text{weight of equal volume of water}}$$

Therefore, when it is said that an electrolyte has a specific gravity of 1.275, it means that, volume for volume, the electrolyte is 1.275 times the weight of pure water.

b. The hydrometer is an instrument for measuring specific gravity. It is used particularly in storage-battery work to measure the specific gravity of the electrolyte and thus indicates the state of charge of the lead-acid cell. The reason why the specific gravity of the electrolyte indicates the state of charge of the cell has already been explained. The hydrometer proper consists of a glass tube 3 to 5 in. long, having a small bulb on one end filled with shot or mercury so that it will float in an upright position when placed in the electrolyte. In accordance with Archimedes' principle which states that *the loss of weight of a body in a liquid is the weight of the displaced liquid*, the hydrometer may be calibrated to show the weight of the liquid compared with the water it displaces. The calibration which shows this value is called the *specific-gravity scale* and appears on the stem of the hydrometer.

In open-jar batteries it is only necessary to float this hydrometer in the electrolyte to take a reading, but for portable use a hydrometer syringe must be used.

c. The hydrometer syringe (see Fig. 419) consists of a glass barrel or tube, a hydrometer which fits inside the glass barrel, a small soft- or hard-rubber tube for reaching the electrolyte, a rubber bulb which fits on one end of the barrel, and a soft-rubber plug for joining the barrel to the tube.

To take a reading, insert the rubber tube in the cell, squeeze the bulb, and then slowly release it, drawing up just enough electrolyte from the cell to freely float the hydrometer, making sure that the stem of the hydrometer does not strike the tube at the bottom of the barrel. The reading on the stem of the hydrometer at the surface of the liquid is the specific-gravity reading of the electrolyte. After testing, the electrolyte must always be returned to the cell from which it was taken.

If a record is kept of the specific-gravity reading of a cell, it is well also to record the temperature and level of the electrolyte, as both of these factors affect the gravity reading to a certain extent.

After adding water, a hydrometer reading should not be taken until the water added has been mixed with the electrolyte by charging the battery until it gasses.

Great care should be taken in handling the acid taken up by the hydrometer syringe, and the syringe hose is best left in the cell while the reading is taken; that is, do not lift up the hydrometer syringe when it is full of electrolyte. Electrolyte has a very destructive effect on clothing and unprotected wood and metals. If accidentally spilled, its effects may be neutralized by *immediately* applying and washing with ammonia or sodium bicarbonate in liberal doses.

d. Temperature Correction.—Since electrolyte, like most substances, expands when heated, its specific gravity is affected by a change in temperature. If electrolyte

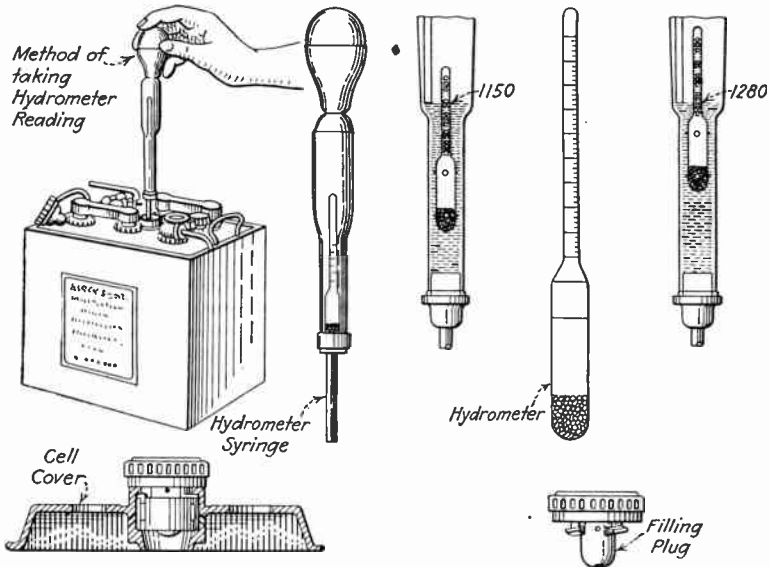


FIG. 419.—Hydrometer.

has a certain specific gravity at a temperature of 80°F. and is heated, the heat will cause the electrolyte to expand and, although the actual strength of the solution will remain the same as before heating, the expansion will cause it to have a lower gravity, a decrease of approximately one point (0.001) for each 3°F. rise in temperature. For example, if electrolyte has a specific gravity of 1.275 at 80°F. and the temperature is raised to 83°F., this increase in temperature will cause the electrolyte to expand and the gravity to drop from 1.275 to 1.274.

On the other hand, if the temperature has been lowered from 80 to 77°F., this would cause the specific gravity to rise from 1.275 to 1.276. Since change of temperature does not alter the strength of the electrolyte, changing its specific gravity only, the gravity reading should be corrected one point for each 3°F. change in temperature.

For 1.275-gravity batteries, 80°F. is considered as normal, while for 1.210 batteries, 70°F. is considered normal. From these points, corrections are made as follows: for every 3°F. rise in temperature, add one point (0.001) to the hydrometer reading, and for every 3° drop below these temperatures, subtract one point (0.001) from the specific-gravity reading. The reason for these corrections will become clear by studying Fig. 420 which is properly labeled to show the reasons for the changes.

395. The voltage of the lead cell is approximately 2 volts on an open circuit, but is higher than this when the battery is being charged, and lower when being discharged. The nominal voltage of a battery is, therefore, the number of cells multiplied by two.

The voltage at any time on discharge or charge depends upon several factors, such as the current rate, the state of charge or discharge, and the temperature. No general averages to cover all conditions can, therefore, be given. The average cell voltage during a 6- to 8-hr. discharge at normal rates is about 1.95 volts with a final voltage of about 1.75 volts. As soon as the cell is put on charge, its voltage rises to about 2.15 volts and then increases during charge until at the end it is between 2.4 and 2.65 volts, depending upon local conditions. The average voltage during entire charge is usually considered 2.33 volts.

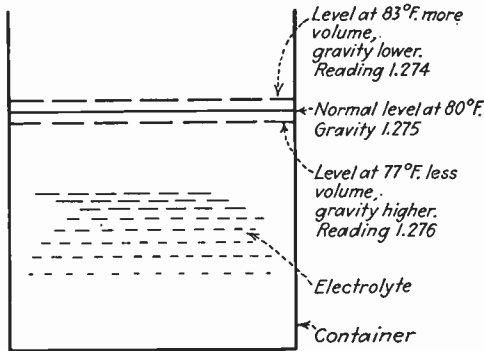


FIG. 420.—Temperature correction for specific-gravity reading.

396. Discharge Rates and Limits. *a. High discharge rates* (amperes) are often confused with overdischarge, a condition resulting when too many ampere-hours are taken out. The effect on the battery of one *as* against the effect of the other is, however, markedly different.

Any properly designed lead cell may be discharged, without injury to the plates, *at any rate of current* that it will deliver. The maximum permissible rate of discharge is limited only by the current-carrying ability (capacity) of the wiring, motor, or other load to which the battery is connected or by the current-carrying ability of the cell terminals and connectors and not by the plates themselves. This means that, if the circuit external to the plates is of sufficient current-carrying capacity, the cells may be actually short-circuited without buckling the plates or injuring the active material.

The number of ampere-hours which may be obtained from a battery is greater for a long, low- or intermittent-rate discharge than for a short, high-rate discharge. This is because the voltage drops faster at the higher rates.

All batteries are given a *normal* capacity rating which is the ampere-hours obtainable under certain working conditions. Therefore, this so-called normal capacity is not the capacity obtainable under all conditions. For example, a battery may be given a normal 8-hr. capacity rating in ampere-hours, but greater capacity can be obtained if the discharge-current rates are made lower. On the other hand, if the rates are above normal, the capacity for such a discharge will be less.

In an emergency, little if any permanent harm will result if the battery is discharged to the full amount that it will give, provided that it is promptly recharged.

It has already been pointed out that the drop in specific gravity of electrolyte should not exceed a certain definite number of points for the particular battery in question.

b. The danger of harm from overdischarge may be illustrated by a comparison between the active material of the plates and the action which goes on when some of the electrolyte is allowed to act on the copper wire at the terminals of a battery. It is noticeable that a comparatively large amount of copper sulphate is formed when only a small quantity of the metal is eaten away by the acid. In the same manner, when acid combines with the lead in the active material, the resulting lead sulphate occupies more space than the active material from which it is formed. The active material of all battery plates is porous, and this expansion of the sulphated material is accommodated by reducing the size of the pores in the active material. All battery plates are designed to accommodate a certain amount of this expansion of the active material during sulphation, and, in batteries of the type under consideration, this is limited to the amount represented by a certain drop in specific gravity.

c. Further discharge, even if it can be obtained at a satisfactory voltage, results in an excessive expansion, which so closes the pores in the active material that it becomes increasingly difficult to recharge the battery properly after an excessive discharge, and, unless a proper recharge is given, the battery is likely to deteriorate. For this reason the discharge of any battery should not be carried beyond the full-discharge point, as indicated by a hydrometer or ampere-hour meter reading.

397. Charging and Charging Rates.—In order to restore a secondary cell (storage battery) to a full-charged condition, it is necessary to reverse the chemical action which occurred during discharge. This may be done by forcing a current through the battery in a direction opposite to the direction of flow of the discharge current. It is evident, therefore, that only a direct current may be used to charge a storage battery. If alternating current only is available at the point where the battery is to be charged, the alternating current must be rectified into a direct current before it can be used for charging purposes.

To cause the charging current to flow in the reverse direction to the discharge current, the positive side of the charging line must be connected to the positive terminal of the battery. The voltage and current values to be used on the charging line will be discussed in Sec. 399, on Charging Methods. These values vary, depending on the job to be done. It is, of course, apparent that the voltage of the charging line must never drop below the voltage of the battery on charge, otherwise the battery would discharge instead of charge. Generally, a charging-line voltage of 2.5 volts per cell is figured for lead batteries.

While a battery is being charged, the amount of sulphate on the plates decreases, and the ability of the plates to give up the acid becomes reduced; in other words, during the early part of a charge the plates can give up the acid at a rapid rate, as there is a large amount of sulphate available. Therefore a battery that is considerably discharged can be charged at a high rate, but, as the charge approaches completion, currents at high rate cannot be utilized, and, if high rates are maintained, only a portion of the current is used to withdraw acid from the plates, and the balance of the current acts to decompose the water in the electrolyte into oxygen and hydrogen, which are given off in the form of gas. Gassing of the battery, therefore, at any time shows whether or not the charging rate is too high. Consequently, when the cells are gassing on charge, the rate of charge should be reduced, so as not to waste the current. Furthermore, the action of the bubbles of gas escaping from the pores of the plates and in "boiling" to the top of the electrolyte have a tendency to wash and wear the active material away from the plates, particularly the positive.

It is a well-known fact that batteries wear out. This wear shows itself to the eye principally in the positive plate, the active material of which softens with use, and, were it not for this unavoidable fact, the life of batteries would be very much longer than at present. As the active material of the plate softens with use, there is a tendency for the softened material on the surface of the plate to fall to the bottom of the jar in the form of sediment. The action of the gas in escaping from the pores of the plates, and the little whirlpools created in the electrolyte when the bubbles of gas boil to the surface, hasten this shedding of material and shorten the life of the battery.

Excessive gassing, therefore, should be avoided if the best life of the battery is to be obtained. A small amount of gassing at low rates and for a short time, at the completion of a charge, is not objectionable, but violent gassing having the appearance of boiling should be avoided.

The value of the charging current at which appreciable gassing begins depends upon several factors, such as state of charge, temperature, specific gravity of electrolyte or type of plate, but the principal factor is the state of charge of the battery. When a battery is fully charged, any rate of charge, however small, will produce gassing, but the rate may be reduced to such a low value that, unless abnormally prolonged, the small amount of gassing which results is practically harmless. This safe rate is called the *finishing* rate. During the earlier stages of the charge, the charging rate may be several times the finishing rate, without producing violent gassing; and the more completely the battery is discharged the higher the charging rate may be without causing excessive gassing.

With the above limitations in mind, it is advisable to charge a battery at relatively low rates if local conditions permit. Of course, the lower the rate of charge the longer the time required to complete the charge. If 8 to 14 hr. are available (the exact time depending upon local conditions), the entire charge may be given at a

low rate, not exceeding the *normal* or *finishing* rate for the particular cell in question. If this length of time is not available, the charging rate should be higher at the start of charge and then reduced to the rate mentioned as soon as the battery begins to gas.

Where an ampere-hour meter is used, the following general arbitrary rule can be given: The charging rate in amperes must be *less than* the ampere-hours still out of the battery until the finishing rate is reached. In other words, the ampere-hour meter will always indicate a rate which should not be *exceeded* at that time. Of course, any rate less than that indicated by the ampere-hour meter will be satisfactory.

The important points are: Avoid gassing and keep the cell temperature below 110°F.

The normal charge and discharge rates are values of current arbitrarily assigned on the basis of economical battery service. As related to discharge, this means the discharge rate at which the cost of watt-hour service is at its lowest for the work in hand. As related to charge, the normal rate is that rate at which the batteries can be most economically charged in commercial service. This applies to all types of lead-acid or Edison storage batteries.

398. Charging Requirements.—The instructions issued by the battery manufacturer for the particular type of battery to be charged should always be carefully studied and followed. Batteries in different classes of service require different methods of charging and are either *cycled* or *float*ed as will be explained later.

a. Marine Emergency Batteries.—The large installations of 60 cells designed to provide emergency power if the ship's generator fails are usually of the portable (1.250 to 1.275 sp. gr.) type and, when not in use, are permanently connected across the ship's line through a high resistance which serves to keep them fully charged and ready for operation at all times. This is the *trickle-charge* method and the battery is said to be *float*ed on the line. It is, of course, necessary to divide the bank of 60 cells into two groups of 30 cells each (by means of a switch) for charging, as shown in Fig. 421, in order that the battery voltage may not exceed the line voltage, which would generally happen if the 60 cells were thrown across the line in series.

If an emergency discharge is necessary, the battery should be immediately recharged at the normal rate (not trickle charged) after which it may be floated on the line as usual.

The trickle-charge or floating-charge system is designed to compensate for local action, which causes the battery to slowly discharge when not in use, and it should keep the battery fully charged at all times after it has once been charged at the normal rate. The floating operation should be checked every week or so, until it is certain that the system gives neither too much nor too little charge. With proper adjustment, the specific gravity of a selected cell, which is always used for the purpose and, therefore, known as the *pilot cell*, will remain practically constant (within 5 to 10 points, if the level of the electrolyte is

kept the same height) and the cells will not be gassing. If the cells gas continually, the battery is receiving too much charge. If the gravity continues to drop, the battery is not receiving enough charge. The charging current may be adjusted, if necessary, by changing the resistance in series with the batteries (usually lamps), using higher wattage lamps to increase the current, or lower wattage lamps to decrease the current, or, if resistance units are used, they may be varied. Another check may be made a week or so later, repeating until it is certain that the system gives neither too much nor too little charge. The adjustment may then be considered correct and will only require occasional checking. The pilot cell should not always be the same cell. An occasional change in the cell selected for this purpose should be made.

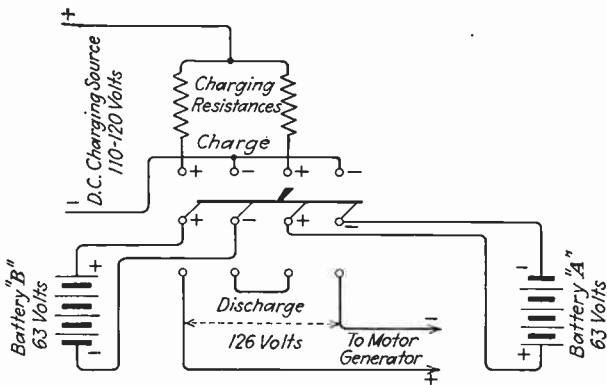


FIG. 421.—Method of charging batteries in parallel banks and discharging in series.

To facilitate the operation of the battery, it is advisable to record three or four times a year the specific gravity of each cell.

Smaller ship-radio battery installations are usually cycle charged, that is, they are not floated across the line but allowed to reach a certain state of discharge after which they are placed on charge at their normal rate.

The *normal rate* is any rate of current which will charge the battery in the shortest possible time without undue gassing. In other words, the normal rate is the maximum current which can be passed through the battery while on charge without excessive gassing.

b. Broadcasting or Other Land-station Batteries.—Batteries in this class of service are usually of the stationary type with a full-charge gravity of 1.210. For plate- and filament-power service the batteries are of the sealed-glass-jar type. In some of the larger stations it has been necessary, because of the large filament loads, to use cells in open lead-lined wood tanks.

Regardless of the type of battery used, they may be cycle charged or floated, depending upon the number of hours they are used each day.

If the battery is used only a few hours each day, and if it carries the load for several hours or the greater part of the day, it is floated.

When a battery is floated across a plate generator, it acts as a filter ironing out any irregularities in the output voltage of the generator, thus making the plate voltage as close to the ideal pure direct-current supply as it is possible to get unless batteries are used solely as the source of power. This is a particularly practical way of getting a constant source of well-filtered direct current for use on speech-input equipment.

399. Charging Methods. *a. Floated Batteries.*—By *floated* batteries, such as are used in some broadcasting and other land stations, are meant those which are continuously connected to the electrical system with which they are used in such a manner that they are normally kept fully or nearly fully charged (except for momentary or emergency discharge) by being constantly maintained at a voltage that will result in a small net charge. The *floating rate* is the sum of the very low current (generally termed *trickle rate*) required to counteract the small internal battery losses plus the average current requirements for the remainder of the circuit. If the latter is zero, the floating rate required becomes the trickle rate for the battery. The required floating or trickle current is automatically provided when the proper voltage is maintained across the battery.

For floated batteries of the 1.210 specific-gravity types, the voltage *directly at the battery terminals* should average very close to 2.15 volts per cell (for example, 129 volts for 60 cells in series) and should be kept between 2.10 and 2.20 volts per cell. If it is continually below 2.10, the charging is insufficient, in which case gravity readings taken from time to time will show a gradual falling off. If continually above 2.20, charging is excessive, in which case unusually frequent addition of water will be required and short life will result. Adjust as necessary until the voltage holds between these limits. Variations outside of these limits, if only of a few minutes' duration, are not harmful and may occur as incidental to normal good operation.

It is very necessary that the voltmeter used be kept in careful calibration, as an error of 5 per cent or even less might have considerable effect upon the reliability and life of the battery.

With the proper voltage across its terminals, the actual current taken by the battery at any time depends on (1) the condition which the battery has had to meet prior to the time under consideration, (2) temperature of the battery, (3) age of the battery. But these variations are in all cases corrective; that is, the battery takes such current as is necessary to keep it in proper condition. This makes it impossible to assign definite values to the current which might be read on an ammeter in the battery circuit, and, therefore, makes voltage operation of the battery much more satisfactory than current operation. However, it may be stated that at normal temperature the normal current flowing to a fully charged battery that has been under constant voltage of 2.15 volts for approximately 1 hr. or more should be between $\frac{1}{4}$ and 1 per cent of the 8-hr. rate of the battery. At higher temperatures or if a discharge has been recently taken (such as a circuit-breaker movement), a current in excess of this will be observed. At lower temperatures, or if the battery has been subject to higher voltage, the observed current will be less or in the latter case may even be temporarily in a discharge direction. If the trickle rate is consistently less than $\frac{1}{4}$ per cent or more than 1 per cent of the 8-hr. rate of the battery, it is recommended that the meters be checked.

Attention is called to the fact that in some cases it is not practicable to connect permanently an ammeter in the battery circuit to indicate the amount of the floating

current, as any high-discharge currents required would, of course, pass through the meter in a reverse direction.

These instructions apply to a battery that is floated continually, 24 hr. daily; but in certain broadcasting stations where the battery is used only when the station is in operation (which may be 18 to 20 hr. daily), the battery is floated on the generator for this length of time at 2.15 volts per cell. In certain installations, duplicate sets of batteries are installed and the sets alternated. One set will carry the load while the other set is on open circuit. Floating the load at 2.15 volts per cell for the 18 to 20 hr. the station is operating will prevent the battery thus floated from discharging, the load being carried on the generator (the primary purpose of the battery in these installations is that they act as filters). But as this voltage is only just high enough to give the battery a small amount of charge, but not enough to bring it up to a state of

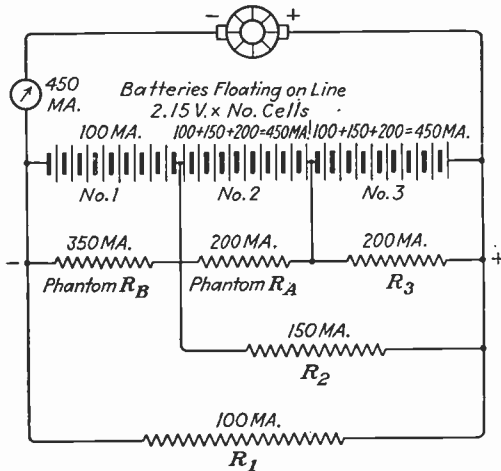


FIG. 422.—Circuit for floating battery on line, utilizing phantom load resistors to equalize discharge.

full charge, an equalizing charge (to be explained later) must be given once a month to bring the battery up to full charge.

Some plate-battery installations require that taps be taken off for different voltages. Under these circumstances the load across the several groups of cells will be different, and therefore some cells will be discharged and charged more than others. To remedy this situation it is necessary to connect phantom-load resistors across all groups except the group handling the largest current drain, and, by means of these resistors, adjust the load across each group so that the value of the load is the same across each group. This is shown by the schematic diagram in Fig. 422. These phantom resistors are taken out of the circuit when the batteries are on open circuit. A voltmeter which can be connected across each section of the battery by means of a plug-and-jack arrangement, or separate voltmeters across each group or section, is advisable in order that the voltage may be held constant across each group at 2.15 volts per cell. When duplicate sets of batteries are provided the sets are changed over from open circuit to floating every 24 hr. As mentioned before, an equalizing charge is then given once a month to bring the cells up to full charge.

b. Equalizing Charge.—An *equalizing* charge is a charge at a rate not higher than the finishing rate for the particular battery, and lower, if practical, and continued until all the cells gas freely and until five consecutive readings of the specific gravity of the pilot cell (or of the lowest gravity cell) show no further increase over a period

of time; this period of time and the interval between the consecutive readings being shown in the table below. The charge may be given by maintaining approximately constant either (1) the charge rate into the battery (constant current method) or (2) the voltage across the battery terminals (constant voltage method).

Interval between readings	Period of time for five readings	Constant-current method charge rate	Constant-voltage method average volts per cell
15 min.	1 hr.	Finishing rate	
30 min.	2 hr.	½ of finishing rate	2.40 volts
1 hr.	4 hr.	¼ of finishing rate	2.35 volts
2 hr.	8 hr.	⅛ of finishing rate	2.30 volts
3 hr.	12 hr.	2.27 volts

In operation which includes frequent charges, the equalizing charge is simply an extension of one of the ordinary charges.

c. Cycled Batteries.—By *cycled* batteries are meant those which are normally allowed to reach a certain state of discharge before being placed on charge, the charge being started manually.

The battery may be charged at any rate in amperes that will not produce gassing or bubbling of the electrolyte or a cell temperature in excess of 110°F. (43°C.). As soon as gassing starts, or before if the temperature reaches this limit, the rate should always be reduced, and the charge should be completed at not higher than the finishing rate. *Do not charge at a higher rate than this while the cells are gassing.* If charging at constant current is more convenient, the entire charge may be given at the finishing rate or at a lower rate.

The best method of charging a battery will depend on the number of cells in the battery, the time available for charging, and the voltage and capacity of the charging apparatus. Wherever possible the charging equipment should be permanently arranged so that the rate of charge is automatically (and not manually) tapered to the finishing rate or to less, by the time the charge is completed. Usually, this is not only possible but very easy and simple to arrange.

Charge the battery at least frequently enough to keep the specific gravity of the electrolyte from falling below the discharged limit.

If the battery has been completely discharged, it should be charged promptly and not allowed to stand completely discharged. Part of a charge is better than none, but it must be remembered that a charge longer than usual is required after a long discharge. Ordinarily too much charge is harmful, but in such a case it is better to give too much charge rather than too little.

The battery should be charged at least once every 6 months.

If the battery has been idle, it should be given a prolonged charge just before using it, continuing until the gravity stops rising.

If the battery requires less than one charge a week, make every charge an equalizing charge.

If the battery requires more than one charge a week, charge until the cells are gassing and until the specific gravity of the pilot cell is within 5 to 10 points of the maximum obtained on the last equalizing charge. *Then stop the charge.* Every sixth or seventh charge should be continued into an equalizing charge.

400. Effect of Temperature on Operation.—The cell temperature should not exceed 110°F The effect of high temperature is primarily

to shorten the life of the wood separators which are installed between the positive and negative plates.

There is always a tendency for wood in contact with sulphuric acid to become carbonized. This tendency is much accelerated at temperatures in excess of 110°F.; and, if a battery is regularly operated at temperatures in excess of this amount, it will probably be necessary to renew separators in the entire battery prior to the time the battery itself ends its useful life.

In the materials used in any commercial storage battery, some impurities are also present which cause very slight action in the cell, even when it is not in active operation. At high temperatures, these internal losses are increased, as is evidenced by the fact that a battery placed in storage will not lose its charge seriously over a period of, say, 6 months if kept in a cool place. If kept at a temperature around 100°F., it will lose much more of its charge in, say, 3 months.

In operation, however, the principal effect of high temperature of a battery is its deteriorating effect on the separators and plates.

In the case of cells in rubber jars if temperatures much in excess of 110°F. exist, there will follow a softening of the sealing compound and a warping of covers of the cells, so that the sealing of the cells will lose its effectiveness. Slopping of the electrolyte will result, leading to deterioration of the wood trays in which the cells are assembled.

Temperature has quite a marked effect on a battery. Low temperature temporarily decreases both the discharge voltage and the ampere-hour capacity which can be taken out of the battery. It is as if the battery were numbed by the cold and unable to make the same effort as at normal temperature. The effect of cold is only temporary, the battery returning to its normal state upon its return to normal temperature even without charge. There is no danger of the electrolyte freezing in a fully charged cell, but in one which is overdischarged or has had the water added without subsequent charging this is likely to occur.

401. Gas and Ventilation.—The hydrogen and oxygen given off from the battery during charge, when unmixed with a large amount of air, form a combination that will explode violently if ignited by an open flame or an electric spark. The battery compartment should, therefore, be open while the battery is being charged, in order that these gases may become mixed with air. Do not bring exposed flame, match, candle, cigar, and so forth, near the battery when charging or shortly after. If the battery is to be charged in a closed room especial attention should be paid to ventilation.

402. Effect of Wrong Charging Polarity.—It is very important that the positive side of the charging line be connected to the positive side of the battery to be charged. It is readily seen that, inasmuch as the object of the charge is to reverse the chemical action which took place

during discharge, the current must flow in the opposite direction to the way it flowed during discharge. If, by accident, the polarity of the charging line is reversed, the chemical action which took place during discharge will be continued, and the effect is to make the negative plates brown and the positive plates gray and to buckle the outside negative plates.

The remedy for a reversed charge is to completely discharge the cells to get rid of the charge in the wrong direction, then short-circuit them as long as current flows (connect a wire across the terminals), and then charge them in the right direction at a reduced rate until the specific gravity reaches a maximum and remains constant for 5 hr. This may require approximately 40 hr., more or less, depending on the type of battery. Under certain conditions the outside negative plate may need straightening. If the battery is operated reversed for any length of time, the negatives will rapidly throw off their material and become useless.

The polarity of the charging line may be determined by a direct-current voltmeter, the terminals of which are marked with the positive (+) and negative (-) signs; the dial pointer will tend to read backward if not properly connected across the line. The positive terminal of the voltmeter must connect to the positive side of the line for the meter to operate properly, and in this way the polarity of the line is indicated.

If a suitable voltmeter is not at hand, the polarity of the line may be determined on circuits of 110 volts or less by dipping the ends of two wires connected to the circuit in a glass of water in which a very small amount (a few pinches) of common table salt, potash, or acid electrolyte has been dissolved. Keep the wires about 1 in. apart, and be very careful not to short-circuit the line. When there is current flowing, gas bubbles will form on both wires, but the wire where the greatest amount of bubbles forms will be the negative side of the line.

Another convenient method of determining polarity is to touch a wet piece of blue litmus paper with two wires from the line, holding them so they touch the paper about $\frac{1}{4}$ in. apart. A red mark will appear at the point where the positive wire touches the paper, if a current is flowing.

Lacking a voltmeter it is perhaps safest to test the polarity of lines carrying voltages between 100 and 550 volts by connecting a neon bulb to the line. This bulb being unidirectional in operation will glow only if, when connected to a direct-current circuit, its anode is connected to the positive side of the line. The polarity of the line becomes apparent from the operation of the tube. A simple line-testing tool using this principle is now on the market.

The polarity of the charging current should be noted before each charge, especially on small installations where no meters are provided to

indicate line condition continuously. This is especially necessary on board ship in the event that the polarity of the ship's generator has been reversed in the engine or dynamo room after a shutdown or change-over.

403. Sulphation and Remedies. *a. Causes.*—Sulphation is such an important phenomenon in connection with the lead storage battery that it is worth while to risk repeating certain points already explained in order to give the matter complete treatment here. During any discharge of a battery, there is being formed sulphate of lead, without which there would be no production of current. If, however, charging is neglected, the sulphate reaches a condition which tends to fill the pores of the plates and make the active material dense and hard. It is this condition which is ordinarily referred to as "sulphated."

To keep a battery in good condition, however, it is neither necessary nor desirable to continue *every* charge until maximum specific gravity has been reached, indicating that no sulphate remains in the plates. A few points less than maximum is sufficient, provided the equalizing charge to maximum specific gravity is given at regular intervals, or, in the case of floated batteries, provided the voltage at the battery terminals is kept within proper limits.

In keeping a battery from becoming sulphated, do not go too far in the other direction and ruin it by overcharging.

The normal lead sulphate formed on discharge of the battery is in a form which absorbs the charge very readily.

When a battery is sulphated, as ordinarily expressed, the sulphate is denser than normal and does not absorb the charge readily, but only with difficulty. When in this condition, the ordinary charge is insufficient.

The cause of this condition is some form of abuse, such as:

Standing discharged.

Habitual undercharging.

Neglecting evidence of trouble in individual cells.

Adding electrolyte or acid to raise the specific gravity, instead of bringing it out of the plates by proper charging.

The active material of sulphated negative plates is generally of light color, and either hard, or granular, gritty, and easily disintegrated. It is the negative plates which require the prolonged charge necessary to restore a sulphated battery.

The last stage of extreme sulphation of negatives is that in which the material deteriorates into what may be described as a mushy condition. This condition almost always results from adding acid to the cells—the "drug-store charge." No treatment is known that will restore cells which have been abused in this manner.

Sulphated positives, unless physically disintegrated or badly buckled, are but little changed in general appearance and can be restored to

operative condition, although their life will not be so great as if they had not been subjected to this abuse.

Sulphated plates should be handled as little as possible.

The user is frequently too ready to conclude that his battery is "sulphated" every time the results obtained are not fully up to his expectations.

If the sediment in a battery has not been allowed to reach the bottom of the plates, and the level of the electrolyte has been properly maintained by restoring level lowered by charging and evaporation with approved water, the battery can be sulphated only because it has not been properly charged or because acid has been added to the electrolyte. An individual cell may become sulphated by external grounding, by an internal short circuit or by drying out as might be caused by failing to add water or to replace a broken jar promptly.

A battery is sulphated only when acid is tied up in the plates in the form of sulphate. When the specific gravity of the electrolyte will rise no higher after continued charging, it shows that there is no more sulphate to be acted upon, since, during charging, the electrolyte receives acid from no other source than the plates.

If a battery is thought to be sulphated, charge the battery fully, and discharge it at the normal rate. If it gives its rated capacity at the battery terminals, the reason for apparent lack of capacity should be looked for elsewhere than in the battery.

If the rated capacity is not obtained, follow the treatment outlined below.

When one or more individual cells have become sulphated while the balance of the battery is in good condition, such cells should be treated separately. It is better to remove them from the circuit.

b. The treatment for removing sulphate is to recharge the battery in the regular manner. When it is considered fully charged, take and record a hydrometer reading of each cell and the temperature of several cells. Charge the battery at a rate as near one-half its normal (or finishing) rate as the charging apparatus will permit. If the temperature reaches 110°F., reduce the current or temporarily interrupt the charge so as not to exceed this temperature.

Hydrometer readings should be recorded at regular intervals (say 3 to 5 hr. apart) to determine if the specific gravity is rising or if it has reached its maximum. Continue the charge, recording the readings until there has been no further rise in any cell during a period of at least ten hours at one-half the normal (or finishing) charge rate. Increase the length of time, proportionately, for lower rates. Maintain the level of the electrolyte at a constant height by adding water after each reading. If water is added just before taking hydrometer readings, the water would not have time to be mixed with the electrolyte by the gassing resulting from charging.

Should the gravity in any cell rise above the normal value, draw off its electrolyte down to the top of the plates and put in water up to the normal level. Continue the charge, and, if the gravity again goes above normal, it shows that acid has been added sometime during the previous operation of the battery. As much electrolyte as possible should then be removed, replaced with water, and the charge continued, repeating the operation as often as necessary.

The treatment may be considered complete only when there has been no rise in the gravity of any cell during a period of at least 10 hr. of continuous charging.

Upon completion of the treatment, the specific gravity of the electrolyte should be adjusted to its normal value, using water or electrolyte, as may be necessary.

By following the few simple rules of operation given in the following section, all the trouble, time and expense of restoring a "sulphated" battery can always be avoided.

c. Water Treatment.—The water treatment is more or less of a drastic measure, to be attempted only as a last resort where prolonged charges in normal gravity are not effective, except in cases where the gravity has been above normal or where the construction of the cell is such that the work involved in lowering the gravity and afterward raising it is not excessive. In cases where prolonged charging in normal gravity would eventually remove the sulphate from the negatives, the water treatment removes it more thoroughly and in much less time.

This treatment does not benefit the positives. In fact, if the charging in reduced gravity is carried to excess, the positive grids may suffer somewhat. Therefore, it should not be attempted in cases where positive grids are cracked, formed, or corroded to any appreciable extent.

The water treatment consists of:

1. Reducing the gravity of the electrolyte until it is between approximately 1.050 and 1.100 specific gravity.
2. Charging at one-half the normal (or finishing) rate to both a 20-hr. gravity and a 20-hr. voltage maximum. If the gravity rises to 1.150, electrolyte should be removed and replaced with water, so that the gravity is between 1.050 and 1.100. Repeat this step until the maximum gravity obtained by charging is below 1.150.
3. Increasing the gravity by adding electrolyte and by charging at the finishing or normal rate until at the end of a 3-hr. gravity maximum the gravity is somewhat lower than normal operating value. The specific gravity of the electrolyte cannot be brought back to normal once a cell has required the water treatment.

404. Care and Operation of Lead Cells.—At various places throughout this chapter are given various explanations and rules to be followed in the operation of lead cells. Some types of cells will require certain kinds of care not required by other types of batteries. Each installation must be studied for its own requirements. The *instructions* supplied by the battery manufacturer for the particular type of battery should be carefully studied and followed.

The following points are important in the operation of any type of lead-battery installation.

- a.* Keep open flames away from the battery at all times during operation.
- b.* Replace spilled electrolyte before charging.
- c.* When water in cell evaporates, add distilled or other approved water.
- d.* Never allow cells to remain in a discharged condition.
- e.* Mix electrolyte in clean earthen or glass jars.
- f.* Allow newly mixed electrolyte to cool before putting into cells.
- g.* Never pour water into acid, injurious action may take place.
- h.* Never allow salt to get into cell.
- i.* Use only chemically pure acid (not, however, concentrated) and water.
- j.* Always provide plenty of ventilation for battery enclosure.

k. If burned by sulphuric acid, apply ammonium hydroxide (*ammonia*) or baking soda (water paste). Dress with aqueous solution of boric acid (half-saturated) to prevent infection.

l. Don't charge at too high rate.

m. Always keep battery charged and ready for operation.

n. When not in use, keep battery floating on line (usually).

o. Take frequent full-load voltage readings.

p. Take occasional hydrometer readings.

q. Keep level of electrolyte to proper height.

r. Give battery an equalizing charge monthly.

s. Keep the batteries clean and their tops dry.

t. Keep all electrical connections clean and bright and free from corrosion.

u. Do not add acid unless some has been lost due to spilling or flooding, and not then unless gravity does not come up with charging.

v. Keep vent plugs in place unless filling cells or taking hydrometer readings.

w. Never add any special powders, liquids, or jellies to the cells to bring back life; these treatments are useless.

x. Avoid undercharging and excessive charging.

y. Keep trays properly wedged.

z. Establish a routine of battery care.

405. Troubles and Their Remedies.—Certain troubles may be encountered in the operation of lead cells and some of the most common of these, in addition to those already treated, will now be covered in detail.

a. *Impurities in the electrolyte* will cause a cell to work irregularly. Should it be known that any impurity has got into a cell, it *should be removed at once*. In case removal is delayed and any considerable amount of foreign matter becomes dissolved in the electrolyte, this solution should be replaced with new immediately, thoroughly flushing the cell with water before putting in the new electrolyte. If in doubt as to whether the electrolyte contains impurities, a sample should be submitted for test.

b. *Sediment* which collects underneath the plates need cause no alarm unless it deposits too rapidly, in which case there is something wrong with the way the battery is operated. In a new battery there is always a thin layer at the start. As the battery wears, the sediment becomes higher, but in batteries which are floated the plates usually wear out before the sediment space is filled.

c. *Excessive Charging.*—Experience shows that one of the most harmful practices from which a radio-battery installation suffers is overcharging. This means continuing the charge after the battery has chemically reached a state of full charge. The effect of this excessive charging is that it causes undue wear on the plates and materially shortens the life of the battery by loosening the active material and depositing it as sediment at the bottom of the jar. The positive plates are especially affected. The only remedy is to stop excessive charging.

d. *Cracked or Broken Jar.*—A slight crack may go unnoticed for a long time, and considerable water will be added before it is noticed that

the electrolyte has been leaking. The plates will not be affected noticeably, but the electrolyte will be weak and will need strengthening by adding acid to replace that lost by the leakage. If the leakage was enough to expose the plates to air and partly dry them, the cell will need some extra charging. Partly dried negative plates always need some excess charge.

The remedy, of course, is to replace the broken jar with a new one. On board ship, however, it is not usually the practice to do this kind of work while underway. In these circumstances the leak might be reduced, or stopped, by applying paraffine, wax, or pitch to the crack. This is, of course, only a temporary measure, and the leaky jar should be replaced as soon as convenient.

If it is not possible to keep electrolyte in the cell on account of the leak, the cell could be disconnected from the bank by cutting the connecting strap with a hacksaw and jumping (not short circuiting) the cell with a heavy-wire conductor. It is well not to expose the plates to air, and for this reason the leak should be minimized as much as possible and leakage replaced by adding distilled water to keep the liquid level above the plates. Ammonia or sodium bicarbonate (not washing soda) may be used to neutralize spilled acid or electrolyte, if applied immediately.

II. THE NICKEL-IRON-ALKALINE BATTERY

406. The nickel-iron-alkaline cell, manufactured only by the Edison Storage Battery Company and known as the *Edison cell*, differs in all respects from the lead-acid cell. The two types differ widely from the standpoint of construction, material used, electrical characteristics and chemical reactions, and have nothing in common.

a. The positive plate consists of perforated nickered-steel tubes, arranged in rows and filled with alternate layers of nickel hydroxide and exceedingly thin flakes of pure nickel.

b. The negative plate consists of a grid of cold-rolled nickered sheet steel holding a number of rectangular pockets filled with iron oxide. The construction and assembly of the plates are shown in Fig. 423.

c. The separators used to separate the plates are narrow strips of especially treated hard rubber which is not injured by the electrolyte. The grid separators are provided with grooves which take the edges of the plates, spacing and insulating them from the steel container.

d. The electrolyte consists of a 21 per cent solution of potassium hydroxide in distilled water to which is added a small percentage of lithium hydrate. The electrolyte has a specific gravity of about 1.200 at 60°F. and does not vary appreciably in density during charge and discharge. Frequent hydrometer readings are therefore unnecessary as they are no indication of the state of charge.

e. *The container* is made of cold-rolled nickled sheet steel. Corrugations provide additional container strength, and the soundness of the top, bottom, and side seams is secured by oxyacetylene welding. Each cell is provided with a filling aperture for flushing. The filling-aperture valve is constructed so that it readily permits the escape of gas when the cell is charging but retards spraying the evaporation of the electrolyte and the entrance of impurities into the cell through the filling aperture.

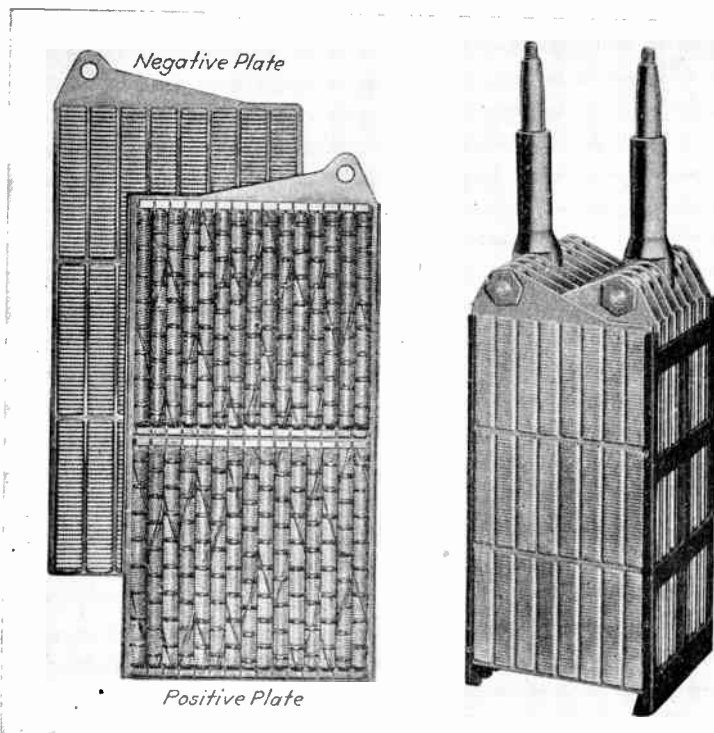


FIG. 423.—Positive and negative plates, and element assembly for Edison battery. (Courtesy of Edison Storage Battery Company.)

f. *Assembly and Connections.*—Nickel-iron-alkaline cells are assembled into batteries in various-sized hardwood trays. Steel suspension bosses, spot-welded to the sides of the containers, fit into hard-rubber buttons recessed into the tray slats. Thus assembled, each cell is supported firmly in place and insulated from the tray and from adjacent cells. This tray assembly is flexible and permits efficient utilization of the space available for battery installation. Connecting and disconnecting cells is a simple mechanical process. Forged nickled-steel, tapered lugs swaged upon heavier copper connecting links are used to connect adjacent cells. The internal taper of the lugs fits the corresponding taper of the pole pieces and insures an excellent electrical connection at these points.

A detailed and labeled view showing the construction of the Edison cell is shown in Fig. 424.

407. Care and Management.—The care required by the Edison cell is limited to such matters as the addition of distilled water to replenish electrolyte losses due to electrolysis, and the renewal of the electrolyte when its specific gravity falls below 1.16 at 60°F. after a full charge.

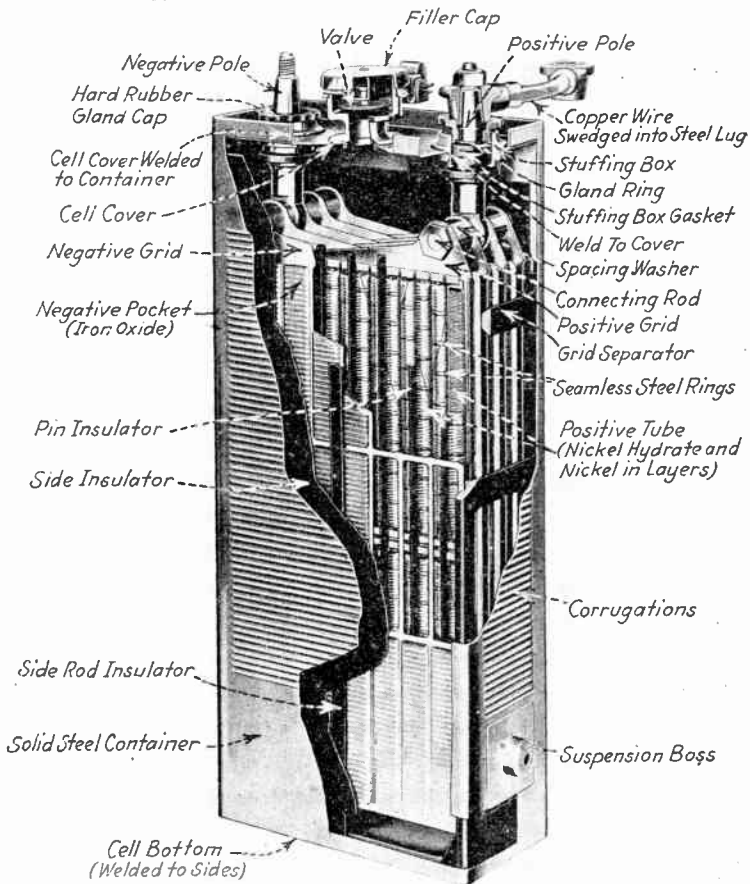


FIG. 421.—Edison cell. (Courtesy of Edison Storage Battery Company.)

The outside of the battery should be kept clean and dry, and the trays and cells given an occasional recoating of Esbalite paint.

408. The Chemical Reaction in an Edison Cell.—The active material in the positive plate of an Edison cell is an oxide of nickel, and that in the negative plate is pure iron. The electrolyte is an alkaline solution consisting of potassium hydroxide with a small quantity of lithium hydroxide in water.

On discharge the pure iron is oxidized. The oxide of nickel in the positive plates is reduced to a lower oxide. On charge the reverse reaction takes place, the oxide of

nickel being raised to a higher oxide and the iron oxide being reduced to pure iron. This transfer of oxygen from the positive to the negative plates on discharge produces electrical energy.

As the potassium hydroxide of the Edison cell electrolyte does not enter into chemical combinations with either the positive or negative active material, the density of the electrolyte does not change appreciably during charge and discharge.

The chemical reactions taking place within an Edison cell on charge and discharge may be indicated as follows:

When Charged

Positive plate is nickel dioxide, NiO_2 .

Negative plate is pure iron, Fe.

Electrolyte is potassium hydroxide and water, $\text{KOH} + \text{H}_2\text{O}$.

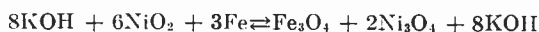
When Discharged

Positive plate is nickel oxide, Ni_3O_4 (partly).

Negative plate is pure iron oxide, Fe_3O_4 .

Electrolyte is potassium hydroxide and water, practically unchanged, $\text{KOH} + \text{H}_2\text{O}$.

The chemical formula for charge and discharge is written:



409. Charging requires direct current. If only alternating current is available, a suitable motor-generator set or other current-rectifying device must be employed.

When discharged to rated capacity, A, B, and C Edison cells require a $6\frac{1}{4}$ -hr. charge at normal rate to recharge them completely.

If rated capacity has not been used on the previous discharge then the charging period may be correspondingly shortened. A nickel-iron-alkaline storage battery may be considered fully charged when its voltage ceases to rise over a period of about 30 min. during charge with constant current flowing.

The positive and negative sides of the charging line should always be connected to the respective positive and negative battery terminals.

Charging the Edison cell in the wrong direction will cause no damage to the cell as long as the temperature of the electrolyte is not allowed to rise above 115°F . The cell will accumulate a slight charge in the reverse direction. If such a charge should be given accidentally to an Edison battery, the battery should be short-circuited to get rid of the reversed charge by discharge, and then the battery should be charged correctly.

Any one of three charging methods may be used:

1. *Constant-current Method.*—This method requires a line voltage of about 1.85 volts per cell to complete the charge and suitable variable resistance in series with the battery so that the charge may be started at approximately 1.54 volts per cell. As the charge progresses, resistance may be cut out and the current thereby maintained constant. The charging rate of A-, B-, and C-type cells should be maintained at about one-fifth of the rated ampere-hour capacity of the battery.

2. *Modified Constant-potential Method.*—This method requires a line voltage of at least 1.84 volts per cell and a fixed resistance in series with the battery equal to (for A-, B-, and C-type cells): $R = \frac{(\text{Line voltage}) - (1.7 \times \text{number of cells})}{\text{Normal rate of battery}}$.

3. *Trickle-charge Method (Low-rate Continuous Charge).*—This method requires a line voltage of from 1.5 to 1.6 volts per cell. The required rate of charge in amperes for the various sizes of A-, B-, and C-type cells is determined as follows:

$$(\text{Rated amp.-hr. capacity} \times 0.16) + (\text{av. amp.-hr. used per 24 hr.} \times 1.10 \div 24)$$

In an emergency, when the time for a normal charge is not available, charging may be done at higher rates than normal. Electrolyte temperatures should not be allowed to exceed 115°F. in the warmest part of the battery during these high-rate or *boosting* charges as they are sometimes called.

410. The state of charge of an Edison battery cannot be determined by hydrometer readings, as there is no change in the specific gravity during charge and discharge. As the voltage of the Edison battery on load when fully charged is 1.37 volts per cell, and as this voltage decreases to 1.00 volt per cell when discharged, the state of charge may be judged by accurate full-load voltage readings. If the battery installation is fitted with an ampere-hour meter, this meter will indicate directly the state of charge of the battery.

411. Ventilation.—As all batteries give off hydrogen gas when gassing, care should be taken to provide sufficient ventilation during charging periods (and also at other times) and to keep away all forms of fire or electrical sparking which might serve to ignite the gas and cause an explosion.

412. Capacity.—The Edison cell maintains its rated capacity over the greater part of its long service life. During the early years of its life a new cell will actually increase in available capacity as much as 10 per cent. Edison cells are guaranteed for varying numbers of years depending upon the class of service for which they are sold. A useful life of from 15 to 20 years and more is not unusual for these batteries in radio service aboard ship.

413. Summary of Data on Lead and Edison Cells.

	Edison cell	Lead cell
Voltage { charged.....	1.37	2.10
	discharged.....	1.75
Specific gravity { charged.....	1.220	1.275
	discharged.....	1.150
Charging voltage per cell.....	1.80-2.00	2.30-2.65
Ampere-hour efficiency, per cent.	80-82	85-95
Watt-hour efficiency, per cent.	60	72-75

III. STORAGE-BATTERY SWITCHBOARDS AND CHARGERS

In order to arrange conveniently the necessary resistances, switches, circuits, circuit breakers, fuses, meters, and any other apparatus necessary to the operation of the battery installation, a control panel or switchboard is used.

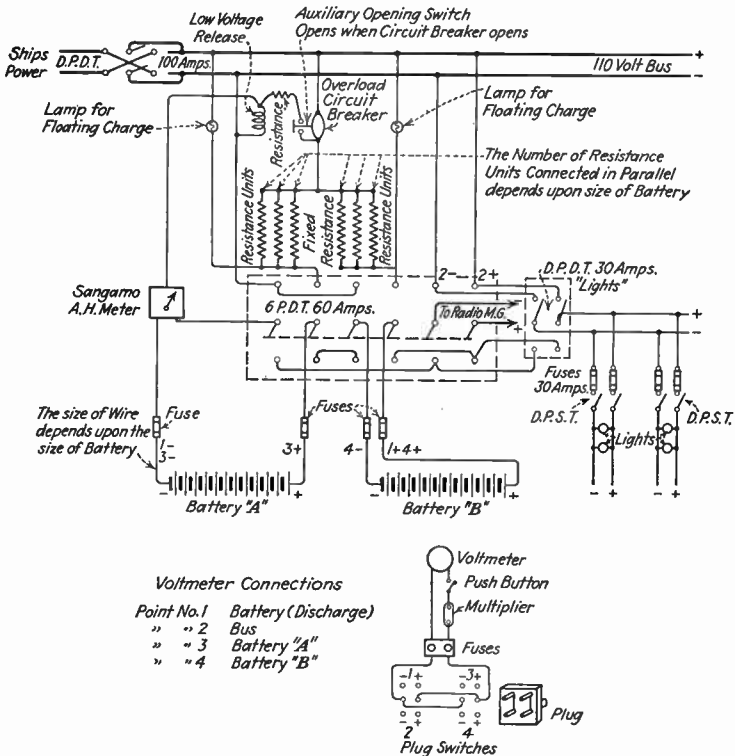


Fig. 425.—Schematic diagram of Exide emergency switchboard.

There are many different types of battery panels, each suited to the particular needs of the installation with which it is used. Some of the types of panels found in commercial and broadcasting service will now be explained.

414. The Exide switchboard,¹ shown in Figs. 425 and 426 is widely used on ships and performs the following functions:

1. Connects batteries in series for discharge.
2. Connects battery into two parallel groups for charge.
3. Provides voltmeter for following readings: (a) ship's-line voltage and polarity, (b) batteries on discharge, (c) group A on charge, (d) group B on charge.
4. Opens or closes ship's-line, or reverses polarity connection thereto.
5. Provides switches for controlling certain emergency lights.

¹ Manufactured by the Electric Storage Battery Co., Philadelphia.

6. Provides ampere-hour meter to indicate state of battery charge.
7. Provides circuit breaker fitted with overload release, low-voltage release, and automatic trip operated by the ampere-hour meter.

The operation of this switchboard is as follows:

NOTE: In these instructions the following abbreviation is used: 6 PDT = six-pole double throw.

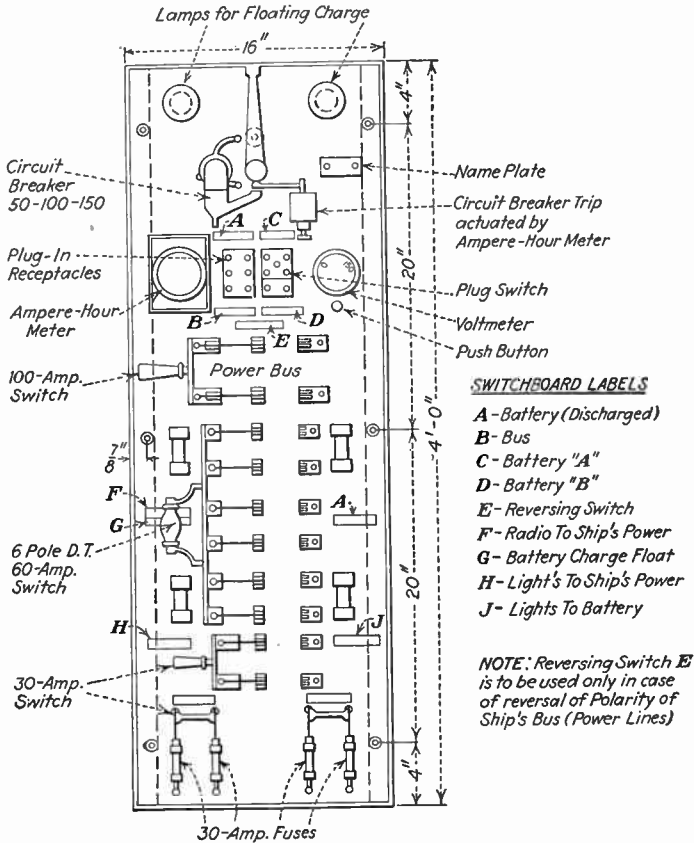


FIG. 426.—Front-panel Exide emergency switchboard.

a. *Checking Polarity.*—First determine that the reversing switch is closed in the proper direction by observing whether the voltmeter reads when the plug switch is in the lower left-hand receptacle which reads the ship's voltage. If it does not read, reverse the reversing switch, then ascertain that the two halves of the battery are also properly connected by taking readings in the upper and lower right-hand receptacles which read "Battery B [and] Battery A and B discharging," respectively.

The voltmeter circuit is normally open and a push-button switch is provided on the switchboard for closing the circuit when it is desired to take a voltage reading. This precaution is taken to prevent inductive effects, incidental to the operation of the radio transmitter, from damaging the meter.

b. *Charging Battery.*—Open the 6 PDT switch. Close the circuit breaker, at the same time holding up the plunger of the low-voltage release coil, and then close the 6 PDT switch to the left. This will place the respective halves of the battery on charge through the charging resistance on the back of the board, which should become uniformly warm. The red pointer on the ampere-hour meter should be set at ampere-hours given in the table of Charging Rates supplied with each installation. The black hand of the ampere-hour meter indicates the state of discharge of the battery at any time. As soon as the charge is started, the black hand will begin to move toward zero and the charge should be complete when it reaches zero. When the black hand reaches zero it makes a contact which opens the circuit breaker by means of the automatic trip, thus automatically cutting off the charge. For the monthly charge, or if for some other reason the battery requires an overcharge, it is necessary to remove the cover from the ampere-hour meter and turn the black hand back, halfway to the red hand. (The ampere-hour meter should be maintained in good operating condition by being overhauled and recalibrated once every 12 to 18 months.)

If the ship's power circuit fails while the battery is charging, the low-voltage release will open the circuit breaker, preventing the battery from discharging back into the bus. The battery can be used for supplying current in such an emergency, even though it may not be fully charged.

c. *Floating Battery.*—With the 6 PDT switch closed to the left and the circuit breaker open, the charging circuit through the resistance units will be open, but the battery will be receiving a floating charge through the two lamps mounted in the upper corners of the switchboard. This is intended to be the normal condition of operation; that is, battery fully charged and floating, with circuit breaker open, and 6PDT switch closed to the left. With the 6PDT switch in this position the radio circuit is connected direct to the bus.

When the battery is floating or charging, the lights cannot be operated from it, and the lower double-pole double-throw switch should then be closed to the left. This throws these emergency lights to the ship's line. The feeder switches for the various light circuits can then be opened or closed, as desired.

d. *Discharging Battery.*—With the circuit breaker open, close the 6 PDT switch to the right. The radio motor generator will now operate from the battery. The emergency lights can also now be operated from the battery, if desired.

e. *Ship's Power Off.*—Whenever the ship's generator is shut down, care should be taken to open the radio-circuit switch on the ship's switchboard, and all switches on the battery switchboard. Do not burn lights from the battery at such times except for emergency.

415. The Edison Battery Switchboard.—This switchboard, shown in Fig. 427, is used in connection with marine installations of the Edison battery. Its function is practically the same as that of the Exide switchboard before described, with a few differences, the most important being the addition of an ammeter on the panel, which may be connected across the various ammeter shunts, shown in the diagram and labeled *Am. Sh.*, by a multi-point selector switch operated by a wheel handle from the front of the panel, and the addition of a heavy-duty rheostat through which the battery may be discharged if it does not get enough action from regular usage.

With new Edison batteries better capacities will result if they are given plenty of work. Likewise, regardless of the age of the battery,

sluggishness or low capacity may result from persistent low-rate discharging or charging, or idleness. For best results these batteries must have constant use, and it is for this reason that a special rheostat is provided through which they may be discharged if they are not used regularly in the normal day's work. The procedure is to discharge completely the battery through the rheostat to zero and then short-circuit it for 2 hr. Follow this by an overcharge which is a charge at the normal rate for a period in excess of the normal hours of charge. This

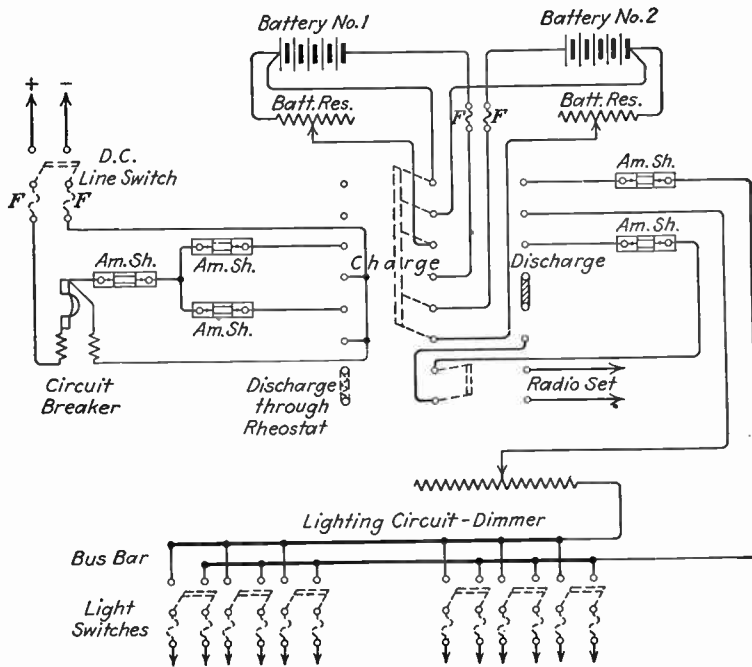


FIG. 427.—Schematic diagram of Edison emergency radio switchboard.

overcharge period varies from 8 to 15 hr., depending on the type of cell employed.

A voltmeter, fuses, rheostat-adjusting handles, circuit breaker, lighting-circuit dimmer, and the switches shown in the schematic diagram are mounted on the front of the panel. The voltmeter, like the ammeter, is controlled by a multi-point selector switch which gives readings on the ship's line, batteries discharging, batteries 1 or 2 on charge or discharge, and the lighting circuit.

416. Charging Panel for Receiver Batteries.—When storage batteries are used for filament supply aboard ship, it is the general practice to provide two batteries so that one may always be available for service while the other one is being charged. A specially designed battery-charging panel supplied for this purpose by the R.M.C.A. and known as

charging panel type UP-858 is shown schematically in Fig. 428. A study of this diagram will make clear how the panel operates. Its installation in the ship's radio room is shown in Fig. 274, in Chapter X on Com-

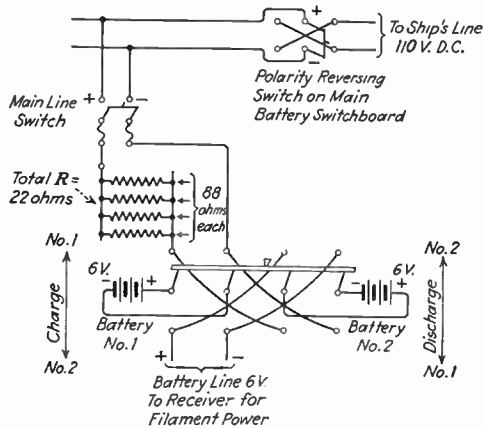


FIG. 428.—Schematic diagram of charging panel for receiver batteries, R.M.C.A. type UP-858.

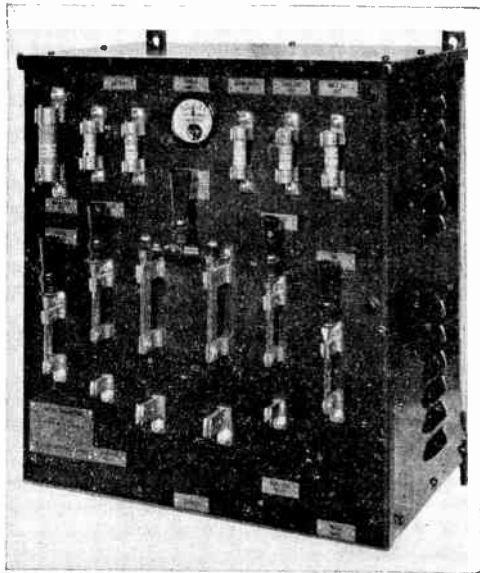


FIG. 429.—Charging panel, R.M.C.A. type for lifeboat equipment. (Courtesy of R.M.C.A.)

munication Transmitters, where the panel is fastened to the bulkhead, to the left of the receiver in the photograph.

417. Lifeboat Panel.—A battery charging panel especially designed for use in lifeboat radio equipments and supplied by the R.M.C.A. is shown

in Fig. 429. A circuit diagram of this panel is shown in Fig. 430. The ampere-hour capacity of the battery required depends upon the size of the lighting load added to the radio load. A typical lighting load on a motor-driven lifeboat may consist of one 250-watt searchlight, one 15-watt Morse light, two 15-watt running lights, and two 15-watt inside lights.

Six cells of Exide TL-17 Ironclad storage battery will operate the radio equipment (type ET-3677 transmitter) and all of the above men-

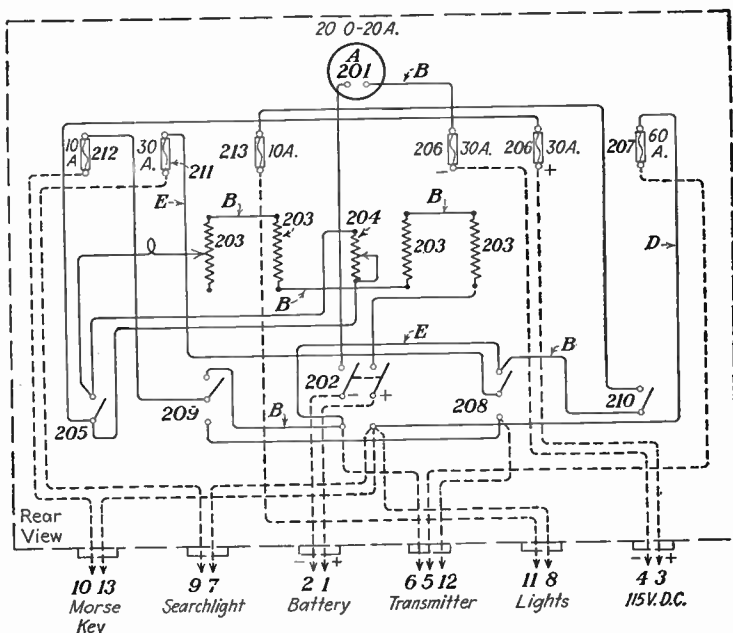


FIG. 430.—Schematic diagram of lifeboat battery panel. (R.M.C.A.)

- | | |
|---|-----------------------------|
| 201. Charging ammeter. | 207. Protective fuse. |
| 202. Charge-discharge line switch. | 208. Searchlight switch. |
| 203. Charging resistors. | 209. Morse light switch. |
| 204. Trickle-charge resistor. | 210. Interior light switch. |
| 205. Trickle-charge-resistor short-circuiting switch. | 211. Searchlight fuse. |
| 206. Line fuses. | 212. Morse light fuse. |
| | 213. Interior light fuse. |

tioned lights continuously for a period of 6 hr. This battery has a capacity of 66.7 amp. for 6 hr.

418. Ampere-hour Meters.—These meters are extensively used with radio storage-battery installations, not only to indicate the state of charge of the battery, but also to terminate the charge automatically when the battery is fully charged. They measure the product of current and time without regard to voltage and thus provide an accurate means of determining the ampere-hours put into, or taken out of, a battery.

The most commonly used ampere-hour meter is the Sangamo type N, a view of which is shown in Fig. 431. This meter is of the mercury-

motor type and comprises a single copper disk rotating in a mercury chamber which is located between the poles of two large permanent magnets. The entire meter is enclosed in a moistureproof steel case which protects it from damage by hard usage. A cross section of the mercury-motor element is shown in Fig. 432.



FIG. 431.—Ampere-hour meter. (Courtesy of Sangamo Electric Company.)

The principles of operation of the mercury-motor type of meter may be explained as follows:

This type of meter employs two permanent magnets between the pole faces of which runs the rotating copper disk as shown in the cross-sectional view in Fig. 432. These permanent magnets furnish both the driving flux and damping, or retarding force. The shape and size of the magnets are such as to allow a very high magnetization. By design the pole tips are close together, being located directly above and below the disk, and produce a very dense and powerful field across the disk. It will be noticed in the cross-sectional view that the current terminal on the left-hand side of the mercury-motor element is labeled as such. A similar current terminal is provided on

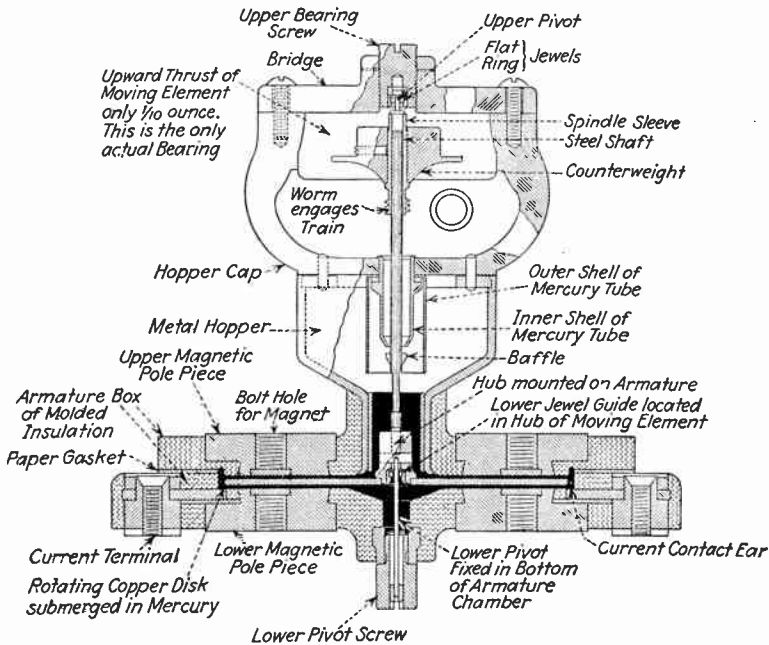


FIG. 432.—Mechanism of ampere-hour meter. (Courtesy of Sangamo Electric Company.)

the opposite side of the motor element, although it is not labeled; instead the current contact ear connecting the current terminal to the mercury is labeled. The

arrangement of current terminal and current contact ear is the same on both the right- and left-hand sides of the drawing. The current to be measured flows from one current terminal to the other, passing first through the one side of the body of high-resistance mercury to the copper disk, through the low-resistance copper disk to its opposite side, and thence through the mercury again to the other current terminal.

As the current flows through the copper disk it rotates owing to the fundamental action by which a torque is produced when a current flows at right angles through a magnetic field. This torque is directly proportional to the product of the current flowing and the magnetic field produced by the permanent magnets, which in this type meter are very powerful and result in a high torque being produced, insuring accuracy on light loads.

A retarding force must be provided to regulate the speed of the disk, in order to make rotation a measurement of the current. Otherwise, the meter would run at a very high rate of speed and would be erratic owing to the effects of friction. As the disk moves in two powerful magnetic fields, eddy currents are generated, which react with the magnetic flux to produce a damping force proportional to the speed of the disk. With these relations, it is evident that the speed of the meter becomes a measure of the current, and a properly geared indicating mechanism (called a *train*), driven from the main shaft, will therefore register ampere-hours.

Compensating Resistor Element.—No storage battery will give out as much energy as is put into it; therefore, a meter designed to control the charging of storage batteries must automatically provide the required excess of charge over discharge. Moreover, this compensating arrangement should be adjustable over a considerable range since it is seldom that two or more battery installations require the same amount of overcharge, which depends on operating conditions.

The resistor, which automatically provides for the excess of charge over discharge, consists of a small auxiliary mercury chamber located in the leakage field of the permanent magnets. This resistor is connected in series with the armature circuit so that a pivoted copper vane within the chamber changes its position when the current through the armature is reversed. In the discharge position the resistor introduces low resistance in the armature circuit and the meter runs at normal speed. On charge, the position of the vane provides a low-resistance shunt path which the current takes in preference to a path of higher resistance in the armature circuit, thus causing the meter to run at a lower speed. The resistor carries practically all the armature current and is consequently very positive in action. It is not shown in Fig. 432.

The state of charge of the battery in ampere-hours is indicated by a hand which revolves over a circular dial. In Fig. 431 this hand is shown pointing to the FULL-CHARGE position. In addition to the movable hand, a red hand, which may be set at any desired point on the circular scale to indicate the discharge point at which the batteries are either manually or automatically placed on charge, is provided. When the movable hand reaches the full-charge position, a contact within the meter actuates a circuit breaker which opens the charging circuit. The batteries are thus protected against the damaging effects of an overcharge.

A circuit breaker is usually furnished with the ampere-hour meter which opens the circuit when its coil is energized owing to contact being made in the meter or elsewhere. This circuit breaker is also made in the no-voltage release type in which an energized coil keeps the armature

holding the contacts in the closed position so long as there is sufficient voltage on the charging line.

419. The Tungar battery charger as developed and manufactured by the General Electric Company is a device for charging storage batteries from an alternating-current line; and it acts as a rectifier and current-limiting device.

The Tungar bulb follows the general principles of operation of the two-element vacuum tube in which a unidirectional electronic flow takes place between the hot cathode and a cold anode. In the bulb there is an inert (argon) gas at low pressure, which is ionized by the electrons emitted by the hot cathode, and which acts as the principal current carrier. The result is that the bulb operates with a very low voltage drop (6 to 8 volts) and is capable of passing a current of several amperes, the current limit depending on the size of the bulb.

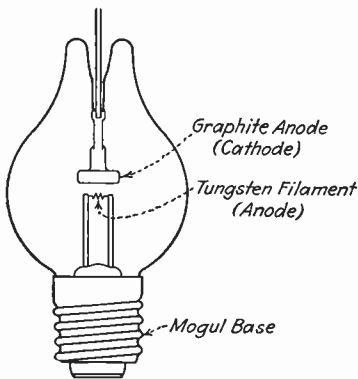


FIG. 433.—Tungar rectifier bulb.

The half-wave bulb consists of a cathode of small tungsten wire coiled into a closely wound spiral to form the filament. The anode is made of graphite and has a relatively large cross section. The bulbs are constructed of high-heat-resisting glass.

The action of the vacuum tube as a rectifier has already been explained in the section of Rectifiers in Sec. 316. The Tungar bulb follows this principle of action, being similar to the mercury-vapor tube in its ionization properties.

All Tungar bulbs are carefully exhausted to the highest possible vacuum and then filled with argon gas in a high state of purity. Certain impurities however, even though present in very small quantities, produce a more or less rapid disintegration of the cathode and also have quite a marked effect on the voltage characteristics of the rectifier. Means must be used to insure absolute purity of the gas from these foreign gases and, to accomplish this, magnesium is introduced into the bulb at the time of manufacture, to react chemically with such impurities as may be present. This reaction keeps the gas in a pure state practically throughout the life of the bulb.

The dark gray or silvery appearance of the bulb is caused by condensation of the purifying agent, magnesium, on the interior of the bulb during manufacture. This is not in the least detrimental to the bulb and does not give any indication of the life of the bulb.

One of the larger sized Tungar bulbs is shown in Fig. 433. This bulb handles 6 amp. of direct current at a pressure of 60 volts. The

recommended filament voltage is 2.2 volts and the approximate filament current 18 amp. It has a maximum inverse-peak voltage of 300 volts. While it is possible to make a Tungar bulb containing two plates for full-wave rectification with one bulb, all, except a small full-wave bulb having a capacity of $\frac{1}{2}$ amp., are of the half-wave type with one anode. Full-wave rectification can, of course, be obtained by means of two half-wave bulbs connected to a transformer with a split secondary. Full- and

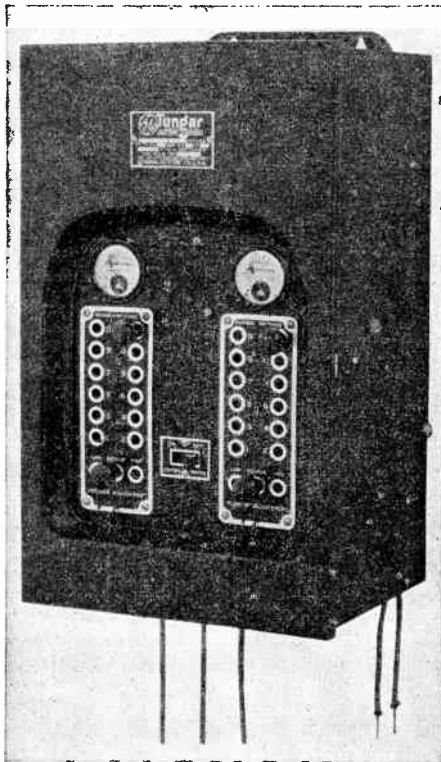


FIG. 434.—Heavy-duty Tungar charger. (Courtesy of General Electric Company.)

half-wave rectifier circuits are shown in Chap. XIII on Rectifier Units and will not be repeated here.

When two half-wave rectifiers are operating from the same supply circuit, it is desirable to reverse the primary connections of one of the outfits, that is, arrange the alternating-current connections so that one outfit operates during one half of the cycle and the other outfit during the second half of the cycle. The effect of this is a considerable increase in the power factor of the circuit.

With rectifiers equipped with insulating transformers it is possible to connect the direct-current terminals to the same load and obtain full-

wave rectification. The primaries must, of course, be reversed as described in the preceding section.

A type of Tungar rectifier used in broadcasting stations is shown in Fig. 434. This is a full-wave charger utilizing two bulbs and designed to charge 24 three-cell lead-storage batteries at 6 amp. or 12 similar batteries at 12 amp.

APPENDIX A

USEFUL TECHNICAL INFORMATION CHARTS AND TABLES

THE NEW TUBE NUMBERING SYSTEM¹

Type numbers for new tubes are now being assigned in accordance with the new system adopted in the early part of 1933 by the Radio Manufacturers Association. A new system was required because practically all of the available two and three digit numbers had been utilized.

The new system, which provides for future expansion of tube types, ordinarily requires only three symbols to identify a tube. These symbols are arranged with a numeral first, then a letter, and finally, a numeral. An example of the new type designation is the 2A5.

New type numbers are formed according to the following simple rules. The first numeral indicates the filament voltage in steps of one volt. For instance, 1 is used for voltages below 2.1; 2 is used for voltages between 2.1 and 2.9 inclusive; 3 for voltages between 3.0 and 3.9, inclusive; et cetera. The digit 1, rather than the digit 2, is used for the 2.0-volt types in order to separate the 2.0- and 2.5-volt tubes. Thus, the 2.0-volt 1A6, and the 2.5-volt 2A5.

The letter is used to distinguish the tube type and is assigned, starting with A, in alphabetical sequence. In the case of rectifiers, however, the assignment is made, starting with Z, in reverse sequence.

The final numeral indicates the number of useful elements *brought out to terminals*. Thus, the 2A5 has five such elements; a heater, a cathode, two grids, and a plate.

While these rules assist to some extent in classifying tubes by filament voltage and function, the significance of the individual symbols will in most cases be inadequate to identify the specific features of a tube.

AUTHORS' NOTE: It is common practice to designate tubes carrying old-style type numbers (such as UX 201-A, CX-301-A, etc.) by using only the last two digits, and letter, if used, preceded by a dash. Thus -01A = UX-201-A or CX-301-A. The reason for this was that the first digit was only a manufacturer's identification; all tubes beginning with 2 being RCA and all those beginning with 3 being made by E. T. Cunningham, Inc. Likewise other manufacturers used different first numbers on their tubes. All tubes whose last two digits were similar had the same characteristics, and were in fact the same tube. Hence the number -01-A refers to a particular type of tube regardless of manufacturer.

This difficulty has now been eliminated by the manufacturers themselves who now drop the first digit and designate these older tubes as RCA-01A or C-01A etc. or simply 01A.

In order to avoid similar confusion we have throughout this book dropped the first digit of the type numbers (on the older tubes), replacing it with a dash. Thus -10 means 210, 310, or 510, etc.

¹ Reprinted from the *Radio Tube Manual*, by courtesy of the RCA Manufacturing Co., Inc.

TABLE X.—THE R.C.A. RADIOTRON-

RCA type	Name	Base S = small M = medium	Socket connections Fig. No.	Cathode type*	Filament or heater		Plate max. volts	Screen, max. volts	Use (Values to right give operating conditions and characteristics for indicated typical use)
					Volts	Amperes			
1A6	Pentagrid converter†	S—6-pin	26	D.-C. filament	2.0	0.06	180	67.5	Converter
1C6	Pentagrid converter†	S—6-pin	26	D.-C. filament	2.0	0.12	180	67.5	Converter
2A3	Power amplifier triode	M—4-pin	1	Filament	2.5	2.5	250 300	Class A amplifier Pushpull-amplifier
2A5	Power amplifier pentode	M—6-pin	15A	Heater	2.5	1.75	250	250	Class A amplifier
2A6	Duplex-diode High- μ triode	S—6-pin	13	Heater	2.5	0.8	250	Triode unit as class A amplifier
2A7	Pentagrid converter†	S—7-pin	20	Heater	2.5	0.8	250	100	Converter
2B7	Duplex-diode pentode	S—7-pin	21	Heater	2.5	0.8	250	125	Pentode unit as r.-f. amplifier Pentode unit as a.-f. amplifier
6A4 also LA	Power amplifier pentode	M—5-pin	6	Filament	6.3	0.3	180	180	Class A amplifier
6A6	Twin-triode amplifier	M—7-pin	24	Heater	6.3	0.8	300	Class B amplifier
6A7	Pentagrid converter†	S—7-pin	20	Heater	6.3	0.3	250	100	Converter
6B7	Duplex-diode pentode	S—7-pin	21	Heater	6.3	0.3	250	125	Pentode unit as r.-f. amplifier Pentode unit as a.-f. amplifier
6C6	Triple-grid detector amplifier	S—6-pin	11	Heater	6.3	0.3	250	100	Screen grid r.-f. amplifier Bias detector
6D6	Triple-grid super-control amplifier	S—6-pin	11	Heater	6.3	0.3	250	100	Screen grid r.-f. amplifier Mixer in superheterodyne
6F7	Triode-pentode	S—7-pin	27	Heater	6.3	0.3	250 250	100 100	Pentode unit as amplifier Pentode unit as mixer
700-A	Detector triode	M—4-pin	1	D.-C. filament	5.0	0.25	45	Grid leak detector
01-A	Detector‡ amplifier	M—4-pin	1	D.-C. filament	5.0	0.25	135	Class A amplifier
10	Power amplifier triode	M—4-pin	1	Filament	7.5	1.25	425	Class A amplifier
11 12	Detector‡ amplifier triode	WD—4-pin M—4-pin	12 1	D.-C. filament	1.1	0.25	135	Class A amplifier
112-A	Detector‡ amplifier triode	M—4-pin	1	D.-C. filament	5.0	0.25	180	Class A amplifier
19	Twin amplifier	S—6-pin	25	D.-C. filament	2.0	0.26	135	Class B amplifier
20	Power amplifier triode	S—4-pin	1	D.-C. filament	3.3	0.132	135	Class A amplifier
22	R.-F. amplifier tetrode	M—4-pin	4	D.-C. filament	3.3	0.132	135	67.5	Screen-grid r.-f. amplifier
24-A	R.-F. amplifier tetrode	M—5-pin	9	Heater	2.5	1.75	275	90	Screen-grid r.-f. amplifier Bias detector

¹ Condensed and reprinted from the *Radio Tube Manual*, by courtesy of the RCA Manufacturing Co., Inc.

* Either a.c. or d.c. may be used on filament or heater, except as specifically noted. For use of d.c. on a.-c. filament types, decrease stated grid volts by one-half (approx.) of filament voltage.

† Grids 3 and 5 are screen. Grid 4 is signal-input control-grid.

‡ For grid-leak detection—plate volts 45, grid return to + filament or to cathode.

CUNNINGHAM RADIO-TUBE CHART¹

Plate supply volts	Grid, volts*	Screen, volts	Screen, ma.	Plate, ma.	A.-C. plate resistance, ohms	Mutual conductance, micro-mhos	Voltage amplification factor	Load for stated power output, ohms	Power output, watts	Type
180	{ - 3.0 min. }	67.5	2.4	1.3	500,000	Anode-grid (No. 2) 135 max. volts, 2.3 ma. Oscillator grid (No. 1) resistor, 50,000 ohms Conversion conductance, 300 micromhos				1A6
180	{ - 3.0 min. }	67.5	2.0	1.5	750,000	Anode Grid (No. 2) 135 max. volts, 3.3 ma. Oscillator Grid (No. 1) Resistor, 50,000 ohms Conversion conductance, 325 micromhos				1C6
250	-4.5	60.0	800	5,250	4.2	2,500	3.5	2A3
300	-62	Self-bias		40.0	Power output is for two tubes at stated load, plate-to-plate			5,000	10.0	2A5
300	-62	Fixed-bias		40.0				3,000	15.0	
250	-16.5	250	6.5	34.0	100,000	2,200	220	7,000	3.0	2A5
250*	- 1.35	0.4	Gain per stage = 50-60				2A6
250	{ - 3.0 min. }	100	2.2	3.5	360,000	Anode grid (No. 2) 200 max. volts, 4.0 ma. Oscillator grid (No. 1) resistor, 50,000 ohms Conversion conductance, 520 micromhos				2A7
100	- 3.0	100	1.7	5.8	300,000	950	285	2B7
250	- 3.0	125	2.3	9.0	650,000	1,125	730	
250§	- 4.5	50	0.65	
100	- 6.5	100	1.6	9.0	83,250	1,200	100	11,000	0.31	6A4 also LA
180	-12.0	180	3.9	22.0	45,500	2,200	100	8,000	1.40	
250	0	Power output value is for one tube at stated load, plate to plate				8,000	8.0	6A6
300	0					10,000	10.0	
250	{ - 3.0 min. }	100	2.2	3.5	360,000	Anode grid (No. 2) 200 max. volts, 4.0 ma. Oscillator grid (No. 1) resistor, 50,000 ohms Conversion conductance, 520 micromhos				6A7
100	- 3.0	100	1.7	5.8	300,000	950	285	6B7
250	- 3.0	125	2.3	9.0	650,000	1,125	730	
250§	- 4.5	50	0.65	
250	- 3.0	100	0.5	2.0	Exceeds 1.5 meg.	1,225	Exceeds 1,500	6C6
250	- 1.95	50	Cathode current 0.65 ma.		Plate coupling resistor 250,000 ohms. Grid coupling resistor 250,000 ohms.§§				6D6
250	{ - 3.0 min. }	100	2.0	8.2	800,000	1,600	1,280	
250	- 10.0	100	Oscillator peak volts = 7.0.				6F7
250	{ - 3.0 min. }	100	1.5	6.5	850,000	1,100	900	
250	-10.0	100	0.6	2.8	Oscillator peak volts = 7.0 Conversion conductance = 300 micromhos				
45	Grid return to (-) filament			1.5	30,000	666	20	'00-A
90	- 4.5	2.5	11,000	725	8.0	01-A
135	- 9.0	3.0	10,000	800	8.0	
350	-31.0	16.0	5,150	1,550	8.0	11,000	0.9	10
425	-39.0	18.0	5,000	1,600	8.0	10,200	1.6	
90	- 4.5	2.5	15,500	425	6.6	11 12
135	-10.5	3.0	15,000	440	6.6	
90	- 4.5	5.0	5,400	1,575	8.5	112-A
180	-13.5	7.7	4,700	1,800	8.5	
135	0	Power output value is for one tube at stated load, plate to plate				10,000	2.1	19
135	- 3.0					10,000	1.9	
90	-16.5	3.0	8,000	415	3.3	9,600	0.045	'20
135	-22.5	6.5	6,300	575	3.3	6,500	0.110	
135	- 1.5	45	0.6¶	1.7	725,000	325	270	22
135	- 1.5	67.5	1.3¶	3.7	325,000	500	160	
180	- 3.0	90	1.7¶	4.0	400,000	1,000	400	24-A
250	- 3.0	90	1.7¶	4.0	600,000	1,050	630	
250**	{ - 5.0 } approx. }	20 to 45	Plate current to be adjusted to 0.1 ma. with no signal					

* Applied through plate coupling resistor of 250,000 ohms.
 § Applied through plate coupling resistor of 200,000 ohms.
 §§ For grid of following tube.
 ¶ Maximum.
 ** Applied through plate coupling register of 250,000 ohms or 500-henry choke shunted by 0.25-megohm resistor.

TABLE X.—THE R.C.A. RADIOTRON-CUNN

RCA type	Name	Base S = small M = medium	Socket connections Fig. No.	Cathode type*	Filament or heater		Plate max. volts	Screen, max. volts	Use (Values to right give operating conditions and characteristics for indicated typical use)
					Volts	Amperes			
26	Amplifier triode	M—4-pin	1	Filament	1.5	1.05	180	Class A amplifier
27	Detector† amplifier triode	M—5-pin	8	Heater	2.5	1.75	275	Class A amplifier Bias detector
30	Detector‡ amplifier triode	S—4-pin	1	D.-C. filament	2.0	0.06	180	Class A amplifier
31	Power amplifier triode	S—4-pin	1	D.-C. filament	2.0	0.13	180	Class A amplifier
32	R.-F. amplifier tetrode	M—4-pin	4	D.-C. filament	2.0	0.06	180	67.5	Screen-grid r.-f. amplifier Bias detector
33	Power amplifier pentode	M—5-pin	6	D.-C. filament	2.0	0.26	135	135	Class A amplifier
34	Super-control r.-f. amplifier pentode	M—4-pin	4A	D.-C. filament	2.0	0.06	180	67.5	Screen-grid r.-f. amplifier
35	Super-control r.-f. amplifier tetrode	M—5-pin	9	Heater	2.5	1.75	275	90	Screen-grid r.-f. amplifier
36	R.-F. amplifier tetrode	S—5-pin	9	Heater	6.3	0.3	250	90	Screen-grid r.-f. amplifier Bias detector
37	Detector‡ amplifier triode	S—5-pin	8	Heater	6.3	0.3	250	Class A amplifier Bias detector
38	Power amplifier pentode	S—5-pin	9A	Heater	6.3	0.3	250	250	Class A amplifier
39-44	Super-control r.-f. amplifier pentode	S—5-pin	9A	Heater	6.3	0.3	250	90	Screen-grid r.-f. amplifier
40	Voltage amplifier triode	M—4-pin	1	D.-C. filament	5.0	0.25	180	Class A amplifier
41	Power-amplifier pentode	S—6-pin	15A	Heater	6.3	0.4	250	250	Class A amplifier
42	Power amplifier pentode	M—6-pin	15A	Heater	6.3	0.7	250	250	Class A amplifier
43	Power amplifier pentode	M—6-pin	15A	Heater	25.0	0.3	135	135	Class A amplifier
45	Power amplifier triode	M—4-pin	1	Filament	2.5	1.5	275	Class A amplifier
46	Dual-grid power amplifier	M—5-pin	7	Filament	2.5	1.75	250 400	Class A amplifier† Class B amplifier‡
47	Power amplifier pentode	M—5-pin	6	Filament	2.5	1.75	250	250	Class A amplifier
48	Power amplifier tetrode	M—6-pin	15	D.-C. heater	30.0	0.4	125	100	Class A amplifier
49	Dual-grid power amplifier	M—5-pin	7	D.-C. filament	2.0	0.120	135 180	Class A amplifier† Class B amplifier‡
50	Power amplifier triode	M—4-pin	1	Filament	7.5	1.25	450	Class A amplifier

* Either a.c. or d.c. may be used on filament or heater, except as specifically noted. For use of d.c. on a.-c. filament types, decrease stated grid volts by one-half (approx.) of filament voltage.

† Grids 3 and 5 are screen. Grid 4 is signal-input control grid.

‡ For grid-leak detection—plate volts 45, grid return to + filament or to cathode.

INGHAM RADIO-TUBE CHART. 1—(Continued)

Plate supply volts	Grid, volts*	Screen, volts	Screen, ma.	Plate, ma.	A.-C. plate resistance, ohms	Mutual conductance, microhms	Voltage amplification factor	Load for stated power output, ohms	Power output, watts	Type	
90	- 7.0	2.9	8,900	935	8.3	26	
180	-14.5	6.2	7,300	1,150	8.3		
135	- 9.0	4.5	9,000	1,000	9.0	27	
250	-21.0	5.2	9,250	975	9.0		
250	-30.0 approx.	Plate current to be adjusted to 0.2 ma. with no signal							
90	- 4.5	2.5	11,000	850	9.3	30	
135	- 9.0	3.0	10,300	900	9.3		
180	-13.5	3.1	10,300	900	9.3		
135	-22.5	8.0	4,100	925	3.8	7,000	0.185	31	
180	-30.0	12.3	3,600	1,050	3.8	5,700	0.375		
135	- 3.0	67.5	0.4 [¶]	1.7	950,000	640	610	32	
180	- 3.0	67.5	0.4 [¶]	1.7	1,200,000	650	780		
180 [§]	- 6.0 approx.	67.5	Plate current to be adjusted to 0.2 ma. with no signal							
135	-13.5	135	3.0	14.5	50,000	1,450	70	7,000	0.7	33	
135	{-3.0} {min.}	67.5	1.0	2.8	600,000	600	360	34	
180		67.5	1.0	2.8	1,000,000	620	620		
180	{-3.0} {min.}	90	2.5 [¶]	6.3	300,000	1,020	305	35	
250		90	2.5 [¶]	6.5	400,000	1,050	420		
100	- 1.5	55	1.8	550,000	850	470	36	
180	- 3.0	90	3.1	500,000	1,050	525		
250	- 3.0	90	1.7 [¶]	3.2	550,000	1,080	595		
100 ^{**}	- 5.0	55	Plate current to be adjusted to 0.1 ma. with no signal							
250 ^{**}	- 8.0	90	2.5	11,500	800	9.2	37	
90	- 6.0	4.3	10,200	900	9.2		
180	-13.5	7.5	8,400	1,100	9.2		
250	-18.0	Plate current to be adjusted to 0.2 ma. with no signal							
90	-10.0	7.0	140,000	875	120	15,000	0.27	38	
100	- 9.0	100	1.2	14.0	115,000	1,050	120	11,600	1.00		
180	-18.0	180	2.4	22.0	100,000	1,200	120	10,000	2.50		
250	-25.0	250	3.8	5.6	375,000	960	360	39-44	
90	{-3.0} {min.}	90	1.4	5.8	750,000	1,000	750		
180		90	1.4	5.8	1,000,000	1,050	1,050		
250	{-3.0} {min.}	90	0.2	150,000	200	30	40	
135 [*]		- 1.5	0.2	150,000	200	30
180 [*]	- 3.0	9.0	103,500	1,450	150	12,000	0.33	41	
100	- 7.0	100	1.6	18.5	81,000	1,850	150	9,000	1.50		
180	-13.5	180	3.0	32.0	68,000	2,200	150	7,600	3.40		
250	-18.0	250	5.5	34.0	100,000	2,200	220	7,000	3.00	42	
95	-15.0	95	4.0	20.0	45,000	2,000	90	4,500	0.90	43	
135	-20.0	135	7.0	34.0	35,000	2,300	80	4,000	2.00		
180	-31.5	180	31.0	1,650	2,125	3.5	2,700	0.82	45	
250	-50.0	250	34.0	1,610	2,175	3.5	3,900	1.60		
275	-56.0	275	36.0	1,700	2,050	3.5	4,600	2.00		
250	-33.0	22.0	2,380	2,350	5.6	6,400	1.25	46	
300	0	Power output values are for two tubes at indicated plate-to-plate load							
400	0	31.0	60,000	2,500	150	5,200	16.0	47	
250	-16.5	250	6.0	47.0	10,000	2,800	28	7,000	2.7		
95	-20.0	95	9.0	50.0	10,000	2,800	28	2,000	1.6	48	
125	-22.5	100	9.0	5.7	4,000	1,125	4.5	2,000	2.5		
135	-20.0	Power output values are for two tubes at indicated plate to plate load							
180	0	11,000	0.17	49	
300	-54.0	Power output values are for two tubes at indicated plate to plate load							
400	-70.0	35.0	2,000	1,900	3.8	4,800	1.6	50	
450	-84.0	55.0	1,800	2,100	3.8	3,670	3.4		
				55.0	1,800	2,100	3.8	4,350	4.6		

[†] Maximum.
^{**} Applied through plate coupling resistor of 250,000 ohms, or 500-henry choke shunted by 0.25-megohm resistor.
^{*§} Applied through plate coupling resistor of 100,000 ohms.
^{*} Applied through plate coupling resistor of 250,000 ohms.
[†] Grid next to plate tied to plate.
[¶] Two grids tied together.

TABLE X.—THE R. C. A. RADIOTRON-CUNN

RCA type	Name	Base S = small M = medium	Socket connections Fig. No.	Cathode type*	Filament or heater		Plate max. volts	Screen, max. volts	Use (Values to right give operating conditions and characteristics for indicated typical use)
					Volts	Amperes			
53	Twin-triode amplifier	M-7-pin§	24	Heater	2.5	2.0	300	Class B amplifier
55	Duplex-diode triode	S-6-pin	13	Heater	2.5	1.0	250	Triode unit as Class A amplifier
56	Super-triode amplifier detector †	S-5-pin	8	Heater	2.5	1.0	250	Class A amplifier Bias detector
57	Triple-grid amplifier detector	S-6-pin	11	Heater	2.5	1.0	250	100	Screen-grid r.-f. amplifier Bias detector
58	Triple-grid super-control amplifier	S-6-pin	11	Heater	2.5	1.0	250	100	Screen-grid r.-f. amplifier Mixer in superheterodyne
59	Triple-grid power amplifier	M-7-pin§	18	Heater	2.5	2.0	250	As triode †† Class A amplifier
							250	250	As pentode § Class A amplifier
							400	As triode Class B amplifier
71-A	Power amplifier triode	M-4-pin	1	Filament	5.0	0.25	180	Class A amplifier
75	Duplex-diode high μ -triode	S-6-pin	13	Heater	6.3	0.3	250	Triode unit as Class A amplifier
76	Super-triode amplifier detector †	S-5-pin	8	Heater	6.3	0.3	250	Class A amplifier Bias detector
77	Triple-grid amplifier detector	S-6-pin	11	Heater	6.3	0.3	250	100	Screen-grid r.-f. amplifier Bias detector
78	Triple-grid super-control amplifier	S-6-pin	11	Heater	6.3	0.3	250	125	Screen-grid r.-f. amplifier
79	Twin-triode amplifier	S-6-pin	19	Heater	6.3	0.6	250	Class B amplifier
85	Duplex-diode triode	S-6-pin	13	Heater	6.3	0.3	250	Triode unit as Class A amplifier
89	Triple-grid power amplifier	S-6-pin	14	Heater	6.3	0.4	250	250	As triode †† Class A amplifier
									As pentode § Class A amplifier
									As triode Class B amplifier
V99 X99	Detector † amplifier triode	S-4-nub S-4-pin	10 1	D.-C. filament	3.3	0.063	90	Class A amplifier

* Either a.c. or d.c. may be used on filament or heater, except as specifically noted. For use of d.c. on a.-c. filament types, decrease grid volts by one-half (approx.) of filament voltage.

† For grid-leak detection—plate volts 45, grid return to + filament or to cathode.

§ Requires different socket from small 7-pin.

|| Grids 1 and 2 connected together. Grid 3 tied to plate.

†† Two grids tied together.

||| Grid 1 is control grid. Grids 2 and 3 tied to plate.

§|| Grid 1 is control grid. Grid 2 is screen. Grid 3 tied to cathode.

INGHAM RADIO-TUBE CHART. 1—(Continued)

Plate supply volts	Grid, volts*	Screen, volts	Screen, ma.	Plate, ma.	A.-C. plate resistance, ohms	Mutual conductance, microhos	Voltage amplification factor	Load for stated power output, ohms	Power output, watts	Type
250 300	0 0	Power output value is for one tube at stated load, plate to plate				8,000 10,000	8.0 10.0	53
135 180 250	-10.5 -13.5 -20.0	3.7 6.0 8.0	11,000 8,500 7,500	750 975 1,100	8.3 8.3 8.3	25,000 20,000 20,000	0.075 0.16 0.35	55
250 250	-13.5 -20.0	5.0	9,500	1,450	13.8	56
250 250	-20.0 - 3.0	Plate current to be adjusted to 0.2 ma. with no signal				57
250	- 1.95	50	Cathodecurrent 0.65 ma.	2.0	1,225 exceeds 1.5 meg.	1,225 exceeds 1,500	57
250	{-3.0} {min. }	100	2.0	8.2	800,000	1,600	1,280	58
250	-10.0	100	Oscillator peak volts = 7.0			
250	-28.0	26.0	2,300	2,600	6.0	5,000	1.25	59
250	-18.0	250	9.0	35.0	40,000	2,500	100	6,000	3.00	59
300 400	0 0	Power output values are for two tubes at indicated plate-to-plate load				4,600 6,000	15.0 20.0	71-A
90 180	-19.0 -43.0	10.0 20.0	2,170 1,750	1,400 1,700	3.0 3.0	3,000 4,800	0.125 0.790	71-A
250*	- 1.35	0.4	Gain per stage = 50-60		75
250	-13.5	5.0	9,500	1,450	13.8	76
100 250	- 1.5 - 3.0	60 100	0.4 0.5	1.7 2.3	850,000 1,500,000	1,100 1,250	715 1,500	77
250	- 1.95	50	Cathode current 0.65 ma.	Plate coupling resistor 250,000 ohms Grid coupling resistor 250,000 ohms§§			
90 180 250 250	{-3.0} {min. }	90 75 100 125	1.3 1.0 1.7 2.6	5.4 4.0 7.0 10.5	315,000 1,000,000 800,000 600,000	1,275 1,100 1,450 1,850	400 1,100 1,160 990	78
180 250	0 0	Power output value is for one tube at stated load, plate-to-plate				7,000 14,000	5.5 8.0	79
135 180 250	-10.5 -13.5 -20.0	3.7 6.0 8.0	11,000 8,500 7,500	750 975 1,100	8.3 8.3 8.3	25,000 20,000 20,000	0.075 0.160 0.350	85
160 180 250	-20.0 -22.5 -31.0	17.0 20.0 32.0	3,300 3,000 2,600	1,425 1,550 1,800	4.7 4.7 4.7	7,000 6,500 5,500	0.300 0.400 0.900	89
100 180 250	-10.0 -18.0 -25.0	100 180 250	1.6 3.0 5.5	9.5 20.0 32.0	104,000 80,000 70,000	1,200 1,550 1,800	125 125 125	10,700 8,000 6,750	0.33 1.50 3.40	89
180	0	Power output values are for 2 tubes operating at indicated plate-to-plate load				13,600 9,400	2.50 3.50	V99 X99
90	- 4.5	2.5	15,500	425	6.6	V99 X99

|| Grids 1 and 2 connected together. Grid 3 tied to plate.
 § Applied through plate coupling resistor of 250,000 ohms.
 §§ For grid of following tube.

Tube Symbols and Bottom Views of Socket Connections¹

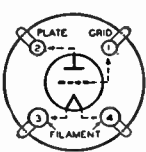


FIG. 1

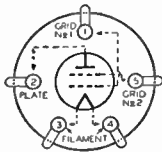


FIG. 7

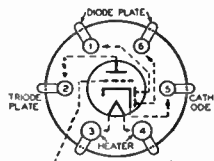


FIG. 13

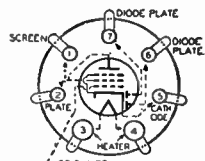


FIG. 21

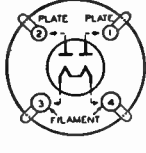


FIG. 2

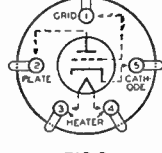


FIG. 8

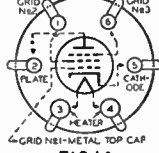


FIG. 14

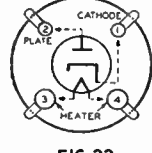


FIG. 22

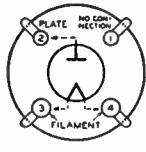


FIG. 3

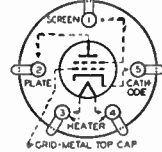


FIG. 9

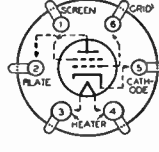


FIG. 15

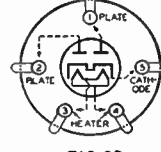


FIG. 23

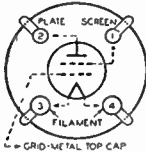


FIG. 4

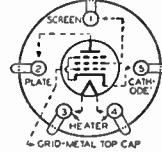


FIG. 9A

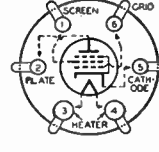


FIG. 15A

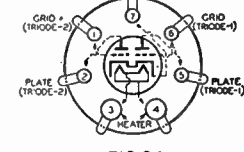


FIG. 24

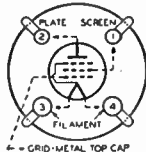


FIG. 4A

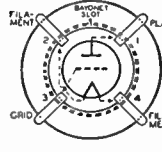


FIG. 10

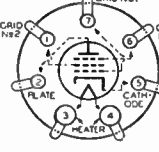


FIG. 18

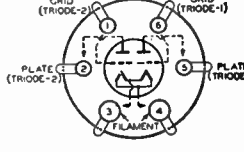


FIG. 25

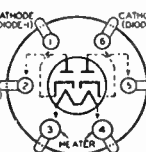


FIG. 5

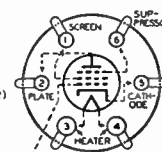


FIG. 11

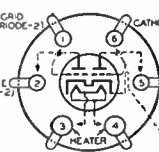


FIG. 19

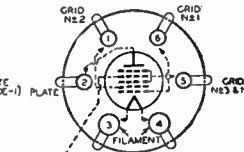


FIG. 26

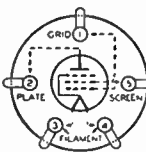


FIG. 6

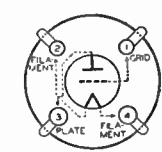


FIG. 12

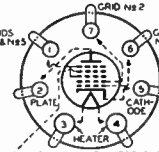


FIG. 20

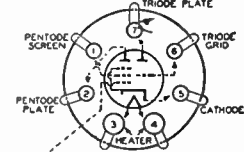


FIG. 27

¹ Reprinted from the *Radio Tube Manual*, by courtesy of the RCA Manufacturing Co., Inc

TABLE XI (PART 1).—CHARACTERISTICS OF TRANSMITTING RADIOTRONS, AMATEUR TYPES
(Condensed from booklet 464, Radiotrons—Amateur Transmitting Types, by courtesy of R.C.A. Radiotron Company, Inc.)

General design		Radiotron number (RCA)													
		203A		211		800		841		852		860		865	
1	Filament or heater voltage.....	10.0 a.c.		10.0 a.c.		7.5 a.c.		7.5 a.c. or d.c.		10.0 a.c.		10.0 a.c.		7.5	
2	Filament current (amp.).....	3.25		3.25		3.25		1.25		3.25		3.25		2.0	
3	Amplification factor.....	25		12		15		30		12		200		150	
4	Grid-plate capacitance (mmf.).....	14.5		15		2.5		7		3		0.08		0.05 max.	
5	Grid-filament capacitance (mmf.).....	6.5		8		2.75		4		2		7.75		10.0	
6	Plate-filament capacitance (mmf.).....	5.5		7		1.0		3		1		7.5		7.5	
<i>Radio-frequency Class B Amplifier*</i>															
7	D.-C. plate voltage (max.).....	1,250		1,250		1,250		450		3,000		3,000		750	
8	D.-C. plate current (max.) (unmodul.) (ma.).....	150		150		45		50		85		85		30	
9	Plate dissipation (max.) (watts).....	100		100		35		15		100		100		15	
10	Screen dissipation (max.) (watts).....											10		3	
11	R.-F. grid current (max.) (amp.).....	6.0		7.5		5.0		5		10		10		5	
12	Typical operation (<i>E_r</i>).....	10.0		10.0		7.5		7.5		10		10		7.5	
13	Plate voltage.....	1,250		1,000		1,000		450		3,000		3,000		750	
14	Screen voltage (approx.).....											300		125	
15	Grid voltage (approx.).....	-45		-80		-55		-12		-250		-55		-30	
16	D.-C. plate current (unmodul.) (ma.).....	105		130		42		36		43		43		22	
17	Peak power output (watts).....	170		160		56		16		160†		160		18††	
18	Carrier output (modul. factor 1.0) (watts).....	42.5		40		14		4		40		40		4.5	
<i>Radio-frequency Class C Amplifier (Oscillator)</i>															
		Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹	Teleph-ony ⁶	Telegraphy ¹¹
19	D.-C. plate voltage (max.) (modul.).....	1,000		1,000		1,000		350		2,000		2,000		500	
20	D.-C. plate voltage (max.) (unmodul.).....	1,250		1,250		1,250		450		3,000		3,000		750	
21	D.-C. plate current (max.) (ma.).....	175		175		80		60		85		100		60	
22	Plate dissipation (max.) (watts).....	67		100		23		35		10		67		15	
23	R.-F. grid current (max.) (amp.).....	6		7.5		6		7.5		4		5		8	
24	D.-C. grid current (max.) (ma.).....	60		60		50		25		25		20		40	
25	Typical operation (<i>E_r</i>) (a.c.).....	10		10		10		7.5		7.5		10		10	
26	D.-C. plate voltage.....	1,000		1,250		1,000		1,250		1,000		1,250		1,000	
27	Screen voltage (approx.).....											2,000		3,000	
28	Grid voltage (approx.).....	-135		-125		-260		-225		-200		-175		-36	
29	D.-C. plate current (unmodul.) (ma.).....	150		150		150		70		70		50		67	
30	D.-C. grid current** (ma.).....	50		25		35		18		15		30		15	
31	Driving power** (watts).....	14		7		14		7		4		1.75		1.25	
32	Power output (watts) (app oximate).....	100		130		100		130		50		65		11.5	

TABLE XI (PART 1).—CHARACTERISTICS OF TRANSMITTING RADIOTRONS, AMATEUR TYPES.—(Continued)

General Design		RCA-250 ²¹	RCA-842	RCA-845	RCA-211	RCA-841	RCA-801		RCA-955 (acorn)	
1	Filament or heater voltage.....	7.5 a.c. or d.c.	7.5 a.c. or d.c.	10.0 a.c.	10.0 a.c.	7.5 a.c. or d.c.	7.5 a.c. or d.c.		6.3	
2	Filament current (amp.).....	1.25	1.25	3.25	3.25	1.25	1.25		0.16	
3	Amplification factor.....	3.8	3	5	12	30	8		25	
4	Grid-plate capacitance (mmf.).....	9	7	15	15	7	6		1.4	
5	Grid-filament capacitance (mmf.).....	5	4	8	8	4	4.5		1.0	
6	Plate-filament capacitance (mmf.).....	3	3	7	7	3	1.5		0.6	
Audio-frequency Class A Power Amplifier and Modulator		Amplifier	Modulator			Resistance coupled, VA.	A.-F. Class A	R.-F. Class B	R.-F. or A.-F. Class A	
7	D.-C. plate voltage (max.).....	450	425	1,250	1,250	425-1,250 ¹⁴	600	600	180	
8	Plate dissipation (max.) (watts).....	25	12	75	75	12	20	20		
9	Typical operation (E_f).....	7.5	7.5	10	10	7.5	7.5	7.5	6.3	
10	D.-C. plate voltage.....	450	425	1,250	1,250	1,000 ¹⁵	600	600	180	
11	Grid voltage ¹² (approx.).....	-84	-100	-209	-80 ²	-9	-55	-75	-5 ^{2a}	
12	Peak grid swing (max.) (volts).....	96			75 ²	9 ²	50			
13	D.-C. plate current (ma.).....	55	28	52	60	2.2	30	45	4.5	
14	Plate resistance (ohms).....	1,800	2,500			40,000	4,300		12,500	
15	Load resistance (ohms).....	4,350		16,000	9,200	250,000	7,800		20,000	
16	Mutual conductance (micromhos).....	2,100	1,200			750	1,840		2,000	
17	Undistorted power output (watts).....	4.6	3.0	24	19.7	225	3.8	(30) ³⁰	135 mw.	
18	Oscillator input per modulator tube (watts).....		14 ¹³							
19	Modulation factor (approx.).....		0.74 ²							
Audio-frequency Class B Power Amplifier and Modulator		RCA-203-A	RCA-800	RCA-852	RCA-211	RCA-841	A.-F. Class B	R.-F. Class C		R.-F. and O Class C
								Telephony ⁹	Telegraphy ¹¹	
20	D.-C. plate voltage (max.).....	1,250	1,250	3,000	1,250	425	600	500	600	180
21	D.-C. plate current ¹⁶ (max.) (ma.).....	175	115	100	175	60	70	60	70	8
22	Plate dissipation ¹⁶ (max.) (watts).....	100	35	100	100	15	20	13.5	20	
23	D.-C. plate input ¹² (max.) (watts).....		85	100		25				
24	Typical operation (2 tubes) (E_f , a.c.).....	10	7.5	10	10	7.5	7.5	7.5	7.5	6.3
25	D.-C. plate voltage (max.).....	1,250	1,250	3,000	1,250	425	600	500	600	180
26	Grid voltage (approx.).....	-45	-70	-250	-100	-5	-75	-190	-150	-35
27	Static plate current (per tube) (ma.).....	13	15	7	10	6.5	4	55	65	7
28	Signal plate current (per tube) (ma.).....	160	65	90	160	60	65	($I_o = 15$ ma.)	($I_o = 15$ ma.)	(1.5; 2 max.) ²³
29	Load resistance (plate to plate) (ohms).....	9,000	21,000	36,000	9,000	7,000	10,000		($I_p = 65$ ma.)	
30	D.-C. plate input (per tube) (watts).....	200	80	90		25	(4.5) ²¹		4	
31	Power output (2 tubes) (watts) (approximate).....	260	106	360	260	28	45	18	25	0.5 ²⁰

List of Legends and Notes pertaining to this table follows Table XIII.

TABLE XI.—(PART 1) CHARACTERISTICS OF TRANSMITTING RADIOTRONS, AMATEUR TYPES.—(Concluded)

General Design		RCA-802 R.-F. Amplifier Pentode			RCA-954 (acorn pentode)		
1	Filament or heater voltage.....	6.3	Note.—Telephony: carrier conditions; for use with a modulation factor up to 1.0. Telegraphy: Key-down conditions.			6.3 a.c. or d.c.	
2	Filament current (amp.).....	0.95				6.3 a.c. or d.c.	
3	Mutual conductance (micromhos).....	2,250				0.5 heater	
4	Amplification factor.....					1400	
5	Grid-plate capacitance (mmf.).....	0.15				2000	
6	Screen-plate capacitance (mmf.).....	0.5				0.007	
7	Input capacitance (mmf.).....	12				3.0	
8	Output capacitance (mmf.).....	8.5				3.0	
Use →		Telephony			Class A Pentode Amplifier	Biased Detector	
		Class B	Class C				Telegraphy
			Suppressor modulation	Grid modulation			
9	D.-C. plate voltage (max.).....	500	500	500	500	250	250
10	D.-C. plate current (ma.) (max.).....	30	30	30	60		
11	D.-C. grid current (ma.) (max.).....		7.5		7.5		
12	D.-C. screen voltage (max.).....	250	200	250	250	100	100
13	D.-C. suppressor voltage (max.).....	40		40	40	100	
14	Plate input (watts) (max.).....				25		
15	Plate dissipation (watts) (max.).....	10	10	10	10		
16	Screen dissipation (watts) (max.).....	4	6	4	6		
17	Typical operation (<i>E_r</i>).....	6.3	6.3	6.3	6.3		
18	D.-C. plate voltage.....	500	500	500	500	250	250
19	D.-C. screen voltage.....	200	200	200	250	100	100
20	D.-C. grid voltage (approx.).....	-28	-90	-130	-100	-3	-6
21	D.-C. suppressor voltage (approx.).....	***	-45	***	40	***	***
22	Internal shield.....	***	***	***	***		
23	Peak a.-f. suppressor volt. (approx.).....		65	50			
24	Peak r.-f. grid voltage (approx.).....	63	125	145	135		
25	D.-C. plate current (ma.).....	25	22	25	45	2	0.1
26	D.-C. screen current (ma.).....	7	28	8	12	0.7	
27	Screen resistor (ohms).....	43,000	10,700	37,500	20,000		
28	D.-C. grid current (ma.).....		4.5	1	2		(Plate load = 250,000 ^{ohms})
29	Driving power (approx.) (watts).....	0.18	0.5	0.8	0.25		
30	Peak power output (watts) (approx.).....	14	14	16	16	(R _p = 1.5 meg-ohms)	
31	Carrier power output (approx.) (watts).....	3.5	3.5	4			
32	Power output (approx.) (watts).....				16		

TABLE XI (PART 2).—CHARACTERISTICS
(Condensed from Technical Information Sheets Supplied by

General design		Radiotron number				
		UV-204A	UV-207	RCA-843 ^P	RCA-844 ^T	RCA-846 ^P
1	Main use.....	0-R.F.	0-R.F.	GP	GP	
2	Number of electrodes.....	3	3	3	4	3
3	Filament or heater (H) voltage.....	11	22	2.5 H	2.5 H	11
4	Current (amp.).....	3.85	52	2.5	2.5	51
5	Type.....	TT	TF	CUC	CUC	TF
6	Average characteristics	E_p 2,000	10,000	425	500	6,500
	I_p, μ, R_p, S_{GP} calculated at $\left\{ \begin{array}{l} E_c \\ E_f \\ E_d \end{array} \right.$	-32	-310	-35	0	-50
		11 a.c.	22 a.c.	2.5	2.5 a.c.	11 a.c.
					180	
7	Plate current (amp.).....	0.125	0.750	0.025	0.013	High Av. Low
8	Amplification factor.....	25	20	7.7	75	44 40 36
9	Plate resistance (ohms).....	6,300	3,500	4,800	125,000	18,500
10	Grid-plate transconductance (microhms).....	4,000	5,700	1,600	600	2,160
11	Approximate direct interelectrode capacities (I.R.E.):					High Av. Low
12	Plate to grid (mmf.).....	17	27	6	.07 ¹	10 9 8
13	Grid to filament (and screen) (mmf.).....	18	18	5	10	7.0 6.5 6.0
14	Plate to filament (and screen) (mmf.).....	3	2	5	8.5	2.0 1.5 1.0
15	Type of cooling.....	Air	Water	Air	Air	Water

Radio-frequency Power Amplifier—Class B

16	Operating d.c. plate voltage (max.).....	2,500	15,000	450	500	7,500
17	Unmodulated d.c. plate (max.) (amp.)..	0.225	1.0	0.030	0.020	0.500
	Plate dissipation (max.) (watts).....	250	10,000	15	15	2,500
18	Screen dissipation (watts).....				3	
19	R.F. grid current (max.) (amp.).....	10	30	2	2	30
20	Cathode to heater potential (max.) (volts)			45		
21	Normal or typical operation at $\left\{ \begin{array}{l} E_p \\ E_c \\ E_f \\ E_d \end{array} \right.$	2,000	13.5 kv.	350		7,000
		-70	-700	-40		-150
		11 a.c.	22 a.c.	2.5		11 a.c.
22	Unmodulated d.c. plate current (amp.)..	0.160	0.900	0.020		0.450
23	Peak output (watts).....	400	16,000	8		4,000 ²
24	Carrier output (watts) (modul. factor 1.0)	100	4,000	2		1,000

Oscillator and Radio-frequency Power Amplifier—Class C

Note 5

Note 5

25	Operating plate potential (max.).....	Telephony	Telegraphy ¹¹			
26	Modulated d.c. (volts).....	2,000		10,000	350	500
27	Unmodulated d.c. (volts).....	2,500	2,500		450	500
28	A.C. (r.m.s.) (volts).....				450	500
29	D.-C. plate current (max.) (amp.).....	0.275	0.275	1.0	0.040	0.030
30	D.-C. grid current (max.) (amp.).....	0.080	0.080	0.200	0.0075	0.005
31	Plate dissipation (max.) (watts).....	250	250	6,600 ⁶	15	15
32	Screen dissipation (max.) (watts).....					3
33	R.-F. grid current (max.) (amp.).....	10	10	20	2	2
34	Normal or typical operation at $\left\{ \begin{array}{l} E_p = \\ E_c = \\ E_f = \\ I_p = \end{array} \right.$	2,000 ⁴	2,000	9,000	350	500
		-175 ²	-175 ²	-2,000	-100 ²	-7
		11	11	22	2.5	2.5 a.c.
		0.250				(E _d = 150)
35	Output (watts).....	350	350	6,000 ²	5	5
36	Socket type (UY) water jacket (UT) mounting (RCA).....	RCA 501, 502	UT-1285	UY	UY plus clip	

TABLE XI (PART 2).—CHARACTERISTICS OF TRANSMITTING RADIOTRONS.—
(Concluded)

Audio-frequency Power Amplifier—Class A

	Radiotron number			
	RCA-843 ^P	RCA-844 ^r	UV-849	UV-851
1 Operating d.c. plate voltage (max.).....	425	425	3,000	2,500
2 Cathode to heater potential (max.) (volts)	45	45		
3 Plate dissipation (max.) (watts).....	12	12	300	600
4 Typical operation at	$\left\{ \begin{array}{l} E_p = 350 \\ E_s = -25 \\ E_f = 2.5 \end{array} \right.$	2,500 -104 11 a.c.	2,000 -65 11
5 D.C. plate current (amp.).....	0.025	0.110	0.270
6 Peak grid swing (volts).....	25	98	60
7 Load impedance (ohms).....	9,500	12,000	3,100
8 Output (5 % second-harmonic) (watts)...	0.95	81	100

Audio Frequency Power Amplifier—Class B

	UV-207	UV-862	UV-849
9 Operating d.c. plate voltage (max.).....	15,000	20,000	3,000
10 Plate input (max.) (watts) ¹⁶	15,000	100,000	825
11 Plate dissipation (max.) (watts) ¹⁶	7,500	50,000	300
12 Typical operation at	$\left\{ \begin{array}{l} E_p = 13,500 \\ E_s = -620^2 \\ E_f = 22 \text{ a.c.} \end{array} \right.$	12,000 -65 33 d.c.	2,500 -130 11 a.c.
13 Signal plate current (zero) (amp.).....	0.20	1.0	0.010 per tube
14 Signal plate current (max.) (amp.).....	1.10	8.5	0.275 per tube
15 Output (watts).....	8,250 peak	50,000 peak	920 nominal

	Modulator		Class A		Class B	
			UV-848	UV-849		
16 Operating d.c. plate voltage (max.).....	12,000		3,000	3,000		
17 Plate dissipation (max.) (watts).....	7,500		300	300		
18 Typical operation at	$\left\{ \begin{array}{l} E_p = 10,000 \\ E_s = -1,040 \\ E_f = 22 \text{ a.c.} \end{array} \right.$		2,500 -104 11 a.c.	2,500 -130 11 a.c.		
Modulation factor.....	0.7					
19 Peak grid swing (volts).....	1.025		98			
20 D.-C. plate current (amp.).....	0.650		0.110	0.350 max.		
21 Oscillator input per modulator tube (watts).....	7,800			825		

TABLE XII.—CHARACTERISTICS OF RECTIFIER TUBES†

Types	Max. plate load d.c., ma.	Max. peak plate current amp.	Max. volts per plate (r.m.s.)	Filament or heater		Max. peak-inverse volts	Remarks
				Volts (a.c.)	Current, amp.		
RCA Receiving:							
5Z3	250	500	5.0	3.0	Full-wave type
12Z3	60	250	12.6	0.3	700	Heater type, half-wave
25Z5	100	125	25.0	0.3	Full-wave, heater type, rectifier doubler
1-v	50	350	6.3	0.3	1,000	Interchangeable with type 1; heater type
80	135	550	5.0	2.0	20-henry input; choke required; full-wave
81	85	700	7.5	1.25		
82	125	0.400	500	2.5	3.0	1,400	Mercury-vapor type; full-wave
83	250	0.800	500	5.0	3.0	1,400	Mercury-vapor type
83v	250	500	5.0	2.0	Heater type full-wave
84/6Z4	50	225	6.3	0.5	Heater type
RCA Transmitting							
214	...	7.5	15,000	22.0	52.0	50,000	Water cooled
217-A	...	0.600	10.0	3.25	3,500	
217-C	...	0.600	10.0	3.25	7,500	
218	...	0.750	11.0	14.75	50,000	No longer specified
219	...	2.5	22.0	24.5	50,000	No longer specified
855	...	5.0	14.5	52.0	50,000	Water cooled; obsolete
857	...	20.0	5.0	30.0	10,000 22,000*	150~ or less 15-50°C. 150~ or less 30-40°C.
							* Forced-air cooling; heater type
866	...	0.600	2.5	5.0	7,500	Mercury-vapor type; tube drop 15 volts
866-A	...	0.600	2.5	5.0	10,000	Mercury-vapor type; tube drop 10 volts
869	...	5.0	5.0	20.0	20,000	Mercury-vapor type; tube drop 15 volts
869-A	...	5.0	5.0	20.0	20,000	Mercury-vapor type; tube drop 10 volts
870	...	450.0	5.0	65.0	7,500	Mercury-vapor type; heater type; forced-air cooled
871	...	0.3	2.5	2.0	5,000	Mercury-vapor type
872	...	2.5	5.0	10.0	7,500	Mercury-vapor type
872-A	...	2.5	5.0	6.75	10,000	Mercury-vapor type; filament transformer 10 amp.
1651	250	4,000	11.0	14.75	No longer specified
W.E. Transmitting:							
214-D and E	150	750	10.0	D 3.0 E 3.25	214-E replaces 214-D interchangeable with 214-D
249-A	...	1.1	2.5	7.0	6,500	Mercury-vapor type
253-A	...	0.500	2.5	3.0	3,500	Mercury-vapor type
258-A	...	1.1	2.5	7.0	6,500	Mercury-vapor type; two-prong base
274-A	150	660	5.0	2.0	Full-wave type
280-A	...	0.5	2.5	3.0	3,500	Mercury-vapor type

TABLE XIII.—CHARACTERISTICS OF WESTERN ELECTRIC VACUUM TUBES

1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
	Normal filament current in amps. (see note 18A) (<i>I</i> = heater)	Normal filament volts (<i>I</i> = heater)	Plate current (ma.) normal unless marked <i>M</i> for maximum	Normal plate voltage 8G-screen grid	Safe plate voltage (max.)	Normal grid voltage (<i>M</i> -max.)	Plate to filament impedance (ohms)	Amplification constant	Output (watts) <i>M</i> -max. continuous; <i>MI</i> -max. intermittent	Plate dissipation (max.) (watts)	Output as an oscillator (watts)	Inter-electrode capacities (mmf.) (see note 17) <i>GK:GP:PK</i>	Use: <i>A</i> -amplifier; <i>D</i> -detector; <i>O</i> -oscillator, <i>M</i> -modulator	Average mutual conductance (micromhos)
101-D ¹⁸	1.0	4.5	15	150	160	- 15	6,500	6.5	0.230 ²¹	3.7:5:2	<i>AOM</i>	
101-F	0.5	4.1	<i>M7.5</i>	190	<i>M-16</i>	5,400	6.5	0.240 ²¹	5.2:5.9:3.7	<i>GP</i>	
102-D ¹⁸	0.95	2.5	1.5	150	160	- 3	100,000	34	<i>DA</i>	
102-E ¹⁸	0.95	2.5	1.5	150	160	- 3	100,000	34	<i>DA</i>	
102-F ¹⁸	0.50	2.3	2.0	150	160	- 3	100,000	33	<i>DA</i>	
102-G ¹⁸	1.0	2.4	1.5	150	160	- 3	100,000	34	<i>DA</i>	
205-D	1.6	5.0	<i>M30</i>	<i>A-300</i>	350	- 20	4,500	7.5	<i>M15</i>	15	5	5.2:4.8:3.3	<i>AOM</i>	2,080
				<i>O-350</i>					<i>MI20</i>					
205-E	1.6	5.0	30	350	350	- 30	4,500	7.5	<i>M15</i>	5	5.2:4.8:3.3	<i>AOM</i>	
									<i>MI20</i>					
211-D ¹⁹	3.0	10.0	65	750	1,000	- 30	4,000	13	<i>M65</i>	100	50	6.3:9.0:3.6	<i>AOM</i>	
211-E	3.0	10.0	3,500	12	<i>MI100</i>	65	6.3:9:3.6	<i>A.-F. A</i>	3,425
212-D ²⁰	6.0	14.0	130	1,500 ²²	2,000	- 60	2,000	17	<i>M200</i>	200	250	19:19:12.4	<i>OM</i>	10,000
									<i>MI250</i>					
214-D & E	Rectifier tube (see Table XII)													
215-A	0.25	1.10	2.0	<i>D-22.5</i>	100	<i>A-9</i>	25,000	6.5	5	1.6:2.6:1.2	<i>ADO</i>	
216-A	1.0	6.0	6.5	<i>A-90</i>	160	- 9	5,900	5.9	<i>AO</i>	
220-B	<i>Wc41</i>	21.5 a.c. or d.c.	1,500	12,000	8,000	40	10,000	24.3:21.5:3.3	<i>AOM</i>	5,000
228-A	<i>Wc41</i>	21.5 a.c. or d.c.	1,500	6,000	2,000	16	5,000	25.9:23.8:3.1	<i>AOM</i>	8,000
231-D	0.060	3.4 d.c.	2.5	135	- 7.5	14,600	7.8	2.4:3.2:2.5	<i>A.-F. A</i>	
232-A	<i>Wc61</i>	20 a.c. or d.c.	<i>M3a.</i>	20,000	7,000	40	25,000	28.9:28.4:0.5	<i>AOM</i>	5,700
236-A	<i>Wc41</i>	21.5 a.c. or d.c.	<i>M2a.</i>	20,000	8,000	40	20,000	29.4:27.8:~.7	<i>AOM</i>	5,000
240-A	<i>Wc41</i>	21.5 a.c. or d.c.	<i>M1.7a.</i>	12,000	9,000	40	10,000	10:26:5	<i>AOM</i>	4,450

242-A	3.25	10.0	M150	1,000	1,250	Class A- 50 B-100 C-150	3,500	12.5	M100	100	M125	6.5:13:4	AOM	6,300
244-A	H1.6	H2.0 a.c. or d.c.	6.0	180	180	- 10	10,000	9.7	3.8:3.3:3.7	DA GP	
247-A	H1.6	H2.0 a.c. or d.c.	3.8	180	- 7	16,000	14.6	3.4:3.2:2.7	DA GP	
249-A	Rectifier tube (see Table XII)	10 a.c.	600	3,000	2,250	10.3	1,000	10:8:6	AOM AOM	4,550
251-A	16.0	10 a.c.	43	450	M-65	1,700	5.0	For fixed grid bias			
252-A ²⁵	2.0	5.0	60	450	M-60	1,500	5.1	For self-biasing grid			6.5:12:4	AOM	2,222
253-A	Rectifier tube (see Table XII)													
●254-A	3.25 ²³	5	60	750 SG 175	80,000	80	20 SG 5	4.6:1:9.4	R.-F. A O	1,000
●254-B	3.25 ²⁴	7.5	M75	750 SG 150	75,000	100	25 SG 5	11.2: .085:5.4	R.-F. A O	1,330
256-A	H1.7	H2.3 a.c.	A triode, contains argon gas at low pressure. Intended for use in special circuits as relay or trigger-action device.											
258-A	Rectifier tube (see Table XII)													
●259-A	H1.6	H2 a.c. or d.c.	7.5	180 SG90M	- 1.5	320,000	480	5.8: .004:14	DA	1,500
261-A	3.25	10	72.5	1,000	- 50	3,500	12	6.5:9:4	GP	6,300
262-A	H0.32	H10 a.c. or d.c.	2.8	180	- 7.5	17,500	14.9	1.8:1:9:4	A.-F. A GP	
264-A	0.3	1.5 d.c.	2.6	100	- 7.0	11,800	7.05	3.5:5.3:2.2	A.-F. A	
268-A	3.25	5.0	5,000	5	30	5.4:2.3:1.1	R.-F. A	1,000
270-A	9.75	10 a.c.	M375	3,000	Class A - 150	1,750	16	350	18.3:21.2:1.8	AOM	9,000
271-A	H2.0	H5.0 a.c. or d.c.	39	400	- 30	2,850	8.5	15	6.5:5.3:3.8	AOM	3,035
272-A	H0.32	H10 a.c. or d.c.	5.9	7,200	5.5	3.4:2.8:2.6	GP	
274-A	Rectifier tube (see Table XII)													
275-A	1.2	5 a.c. or d.c.	52	250	- 60	1,000	2.85	6.8:12:3.2	A.-F. PA	
276-A	3.0	10 a.c.	M125	1,250	Class A- 50 B-100 C-150	3,500	12	Class A 10 B100 C100P	85 100 100	M100	6:9:4	AOM	6,000
277-A	H2	H5 a.c.	A triode, contains argon gas at low pressure. Intended for use in special circuits as relay or trigger-action device.											
279-A	21	10 a.c.	800	3,000	- 275	1,800	10	1,200	15.4:17.9:7.8	AOM	5,550
280-A	Rectifier tube (see Table XII)													
●282-A	3.0 ²⁵	10	M100	1,000 SG 250	70,000	100	70 SG-5M	12.2:0.2:6.8	R.-F. A O	1,430
284-A	3.25	10 a.c.	M150	1,250	1,900	4.7	Class A 52, 5 B-C P 100	85 100	7:8.2:7.8	AOM	2,475
287-A	7.0	2.5 a.c. or d.c.	A triode, contains argon gas at low pressure. Intended for use in special circuits as relay or trigger-action device.											
305-A	3.0	10TT	M100	M1,000	SG250	40,000	56	60(SG5)	10.4:0.14:5.3	R.-F.A.	1,400
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15

LEGENDS AND NOTES FOR TABLES XI, XII AND XIII.

- * As tubes are used under many widely different conditions, these figures should not be used for design purposes without confirmation from the manufacturer.
- ** Subject to wide variations depending on the impedance of the load circuit.
- *** Connected to cathode at socket.
- Tetrode.
- † Plate dissipation 90 watts.
- †† Plate dissipation 12 watts.
- ‡ All rectifier tubes listed in Table XII are *half-wave* type except those labelled *full-wave*.
- A Amplifier.
- A. F. Audio frequency.
- CUC Coated unipotential cathode heater.
- D Detector.
- GP General purpose.
- O Oscillator
- P Preliminary characteristics.
- PA Power amplifier.
- PM Plate modulator.
- R.-F. Radio-frequency (power amplifier).
- SG Screen grid.
- T Tentative characteristics.
- TF Tungsten filament.
- TT Thoriated tungsten filament.
- VA Voltage amplifier.
- W_c Water cooled.
- 1 Filament and screen grounded.
- 2 Approximate.
- 3 And compressed air.
- 4 Unmodulated.
- 5 Plate modulated oscillator and r.-f. power amplifier, class C, telephony. Carrier conditions, modulation factor = 1.0.
- 6 Carrier.
- 7 Although the UV-848 may be used for this service, other radiotrons are recommended as more suitable.
- 8 If plate modulation is used, divide this figure by (1 + modulation factor) to get the maximum d.-c. plate current.
- 9 Plated modulated r.-f. power amplifier, Class B or C, telephony. Carrier conditions; modulation factor = 1.0.
- 10 Socket plus clip.
- 11 Key down conditions.
- 12 Measured from mid-point of a.-c.-operated filament.
- 13 The above data are for an oscillator (or r.-f. amplifier) plate current of 40 ma. at 350 volts.
- 14 The first value (425) is the effective voltage at the plate, which is less than the supply voltage, the second value (1,250), by an amount equal to the voltage drop in the load resistance.
- 15 Plate-supply voltage.
- 16 Averaged over any A.-F. cycle.
- 17 In the Western Electric Tube Chart inter-electrode capacities for triodes are given in the following order: a colon separating each value;

- grid to filament; grid to plate; plate to filament. For screen-grid tubes the values are given as follows: control grid to (heater), cathode and screen grid; control grid to plate; plate to heater, cathode, and screen grid.
- 18 Electrical values given apply when tube is used as amplifier. When used as a detector or for other purposes, the applied voltages may be varied according to circuit conditions and thus cause a corresponding variation in other electrical values.
- 18A An increase of 10 per cent in the filament current may cause as much as 50 per cent decrease in the useful life. For maximum life, therefore, the current should be kept as low as possible to secure the desired output.
- 19 This tube replaces the 211-A (except in 10-A amplifier) and is interchangeable with it, except that it operates at a lower filament current. The discoloration of the tube is due to the manufacturing process and has no effect on the operation of the tube.
- 20 This tube replaces the 212-A tube and is interchangeable with it except that it operates on a lower filament current. The discoloration is due to the manufacturing process and has no effect on the operation of the tube.
- 21 Input in peak values is equal to grid voltage.
- 22 Maximum instantaneous filament-plate peak voltage, 3,000.
- 23 Average thermionic emission 0.6 amp.
- 24 Average thermionic emission 1.0 amp.
- 25 Average thermionic emission 1.25 amp.
- 26 Plate dissipation for 252A is 30 watts.
- 27 Not now listed as an amateur transmitting tube.
- 28 D.-c. resistance in grid circuit not to exceed 0.5 megohm.
- 29 At five meters; moderate reduction down to one meter; below one meter power decreases as wave length decreases.
- 30 This value is the peak power output. Carrier power output approximately one quarter of this value.
- 31 Driving power.
- 32 D.-c. grid current.

TABLE XIV.—WAVE-LENGTH-FREQUENCY CONVERSION TABLE (VERY HIGH FREQUENCIES)

λ		mc./s.	λ m.	mc./s.	λ m.	mc./s.
cm.	m.					
10	0.1	3,000	4.1	73.1	8.1	37.0
20	0.2	1,500	4.2	71.4	8.2	36.6
30	0.3	1,000	4.3	69.7	8.3	36.1
40	0.4	745	4.4	68.1	8.4	35.7
50	0.5	600	4.5	66.6	8.5	35.3
60	0.6	500	4.6	65.2	8.6	34.7
70	0.7	429	4.7	63.8	8.7	34.5
80	0.8	375	4.8	62.5	8.8	34.1
90	0.9	333	4.9	61.2	8.9	33.7
	1.0	300	5.0	60.0	9.0	33.3
	1.1	273	5.1	58.8	9.1	33.0
	1.2	250	5.2	57.7	9.2	32.6
	1.3	231	5.3	56.6	9.3	32.2
	1.4	214	5.4	55.5	9.4	31.9
	1.5	200	5.5	54.5	9.5	31.6
	1.6	187	5.6	53.5	9.6	31.2
	1.7	176	5.7	52.6	9.7	31.0
	1.8	167	5.8	51.7	9.8	30.4
	1.9	158	5.9	50.8	9.9	30.1
	2.0	150	6.0	50.0	10.0	30.0
	2.1	143	6.1	49.2	10.1	29.7
	2.2	136	6.2	48.4	10.2	29.4
	2.3	130	6.3	47.6	10.3	29.1
	2.4	125	6.4	46.9	10.4	28.8
	2.5	120	6.5	46.1	10.5	28.6
	2.6	115	6.6	45.4	10.6	28.3
	2.7	111	6.7	44.8	10.7	28.0
	2.8	107	6.8	44.1	10.8	27.8
	2.9	103	6.9	43.5	10.9	27.5
	3.0	100	7.0	42.9	11.0	27.3
	3.1	96.7	7.1	42.2	11.1	27.0
	3.2	93.7	7.2	41.6	11.2	26.8
	3.3	90.9	7.3	41.1	11.3	26.5
	3.4	88.2	7.4	40.5	11.4	26.3
	3.5	85.7	7.5	40.0	11.5	26.1
	3.6	83.3	7.6	39.5	11.6	25.9
	3.7	81.0	7.7	38.9	11.7	25.6
	3.8	78.9	7.8	38.4	11.8	25.4
	3.9	76.9	7.9	38.0	11.9	25.2
	4.0	75.0	8.0	37.5	12.0	25.0

NOTE: cm. = centimeters; m. = meters; mc/s. = megacycles.

TABLE XV.—RELATION OF WAVE LENGTH (λ), FREQUENCY (f), $2\pi f = (\omega)$, AND $L \times C$ (Reprinted from "The Principles Underlying Radio Communication" by permission of the War Department, Office of the Chief Signal Officer)

λ = wave length in meters.
 f = frequency, cycles per second. (Multiply values in table by 1,000.)
 $\omega = 2\pi$ times the frequency (= $6.28f$). (Multiply values in table by 1,000.)
 L = inductance in centimeters. (1,000 cm. of inductance = 1 microhenry).
 C = capacity in microfarads.

Example 1: Find the capacity which must be connected in series with an inductance of 500,000 cm. (500 μ h.) in order to tune to 2,500 meters. In the table opposite 2,500 meters we find for LC a value of 1,760, which divided by 500,000 yields a quotient of 0.00352. The capacity which must be used is, therefore, 0.00352 mf.

Example 2: What inductance must be placed in series with a condenser of 0.005 mf. in order that the circuit may have a wave length of 600 meters? In the table opposite 600 meters we find for LC a value of 101.4, which divided by 0.005 yields a quotient of 20,280. The inductance which must be used is, therefore, 20,280 cm., or 20.28 μ h.

Example 3: Find the wave length of a circuit having a capacity of 0.0001 mf., and an inductance of 101,600 cm. (101.6 μ h.). The product of the inductance and the capacity is $101,600 \times 0.0001 = 10.16$. In the column headed LC we find that a value of $LC = 10.16$ corresponds to a wave length of 190 meters.

λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm. C in μ f.	λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm.* C in μ f.
1	300,000	1,884,000	0.0003	100	3,000	18,840	2.816
2	150,000	942,000	0.0011	105	2,857	17,940	3.105
3	100,000	628,000	0.0018	110	2,727	17,130	3.404
4	75,000	471,000	0.0045	115	2,609	16,380	3.721
5	60,000	377,000	0.0057	120	2,500	15,710	4.05
6	50,000	314,200	0.0101	125	2,400	15,070	4.40
7	42,900	269,000	0.0138	130	2,308	14,480	4.76
8	37,500	235,500	0.0180	135	2,222	13,950	5.13
9	33,330	209,400	0.0228	140	2,144	13,450	5.52
10	30,000	188,400	0.0282	145	2,069	12,980	5.92
15	20,000	125,600	0.0635				
20	15,000	94,200	0.1129	150	2,000	12,560	6.34
25	12,000	75,400	0.1755	155	1,935	12,150	6.76
30	10,000	62,800	0.2530	160	1,875	11,770	7.20
35	8,570	53,800	0.3446	165	1,818	11,410	7.66
40	7,500	47,100	0.450	170	1,765	11,080	8.13
45	6,670	41,900	0.570	175	1,714	10,760	8.62
				180	1,667	10,470	9.12
50	6,000	37,700	0.704	185	1,622	10,180	9.63
55	5,450	34,220	0.852	190	1,579	9,910	10.16
60	5,000	31,420	1.014	195	1,538	9,660	10.71
65	4,620	28,970	1.188				
70	4,290	26,900	1.378	200	1,500	9,420	11.20
75	4,000	25,120	1.583	205	1,463	9,190	11.83
80	3,750	23,520	1.801	210	1,429	8,970	12.41
85	3,529	22,120	2.034	215	1,395	8,760	13.01
90	3,333	20,920	2.280	220	1,364	8,560	13.62
95	3,158	19,830	2.541	225	1,333	8,370	14.25

* If L is in microhenries, divide the value in the column by 1,000; if in henries, divide by 10^9 .

TABLE XV.—RELATION OF WAVE LENGTH (λ), FREQUENCY (f), $2\pi f = (\omega)$, AND $L \times C$.
 (Continued)

λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm. C in $\mu f.$	λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm. C in $\mu f.$
230	1,304	8,190	14.89	640	469	2,942	115.4
235	1,277	8,020	15.55	650	462	2,896	118.8
240	1,250	7,850	16.22	660	455	2,852	122.5
245	1,225	7,690	16.90	670	448	2,810	126.3
				680	441	2,768	130.2
250	1,200	7,540	17.60	690	435	2,730	134.1
255	1,177	7,390	18.31				
260	1,154	7,250	19.03	700	429	2,692	137.8
265	1,132	7,110	19.77	710	423	2,654	141.9
270	1,111	6,980	20.52	720	417	2,616	145.9
275	1,091	6,860	21.29	730	411	2,580	150.0
280	1,071	6,740	22.07	740	405	2,544	154.0
285	1,053	6,620	22.87	750	400	2,510	158.3
290	1,035	6,500	23.66	760	394.8	2,476	162.6
295	1,017	6,380	24.50	770	389.6	2,443	166.8
				780	384.6	2,412	171.4
300	1,000	6,280	25.33	790	379.8	2,382	175.6
310	968	6,080	27.05				
320	938	5,890	28.83	800	375.0	2,353	180.1
330	909	5,700	30.66	810	370.4	2,325	184.7
340	882	5,540	32.55	820	365.9	2,297	189.3
350	857	5,380	34.48	830	361.4	2,270	194.0
360	833	5,230	36.48	840	357.1	2,242	198.5
370	811	5,090	38.54	850	352.9	2,214	203.4
380	790	4,963	40.7	860	348.8	2,188	208.2
390	769	4,830	42.8	870	344.8	2,162	213.2
				880	340.9	2,138	217.9
400	750	4,710	45.0	890	337.1	2,115	222.9
410	732	4,590	47.3				
420	714	4,480	49.7	900	333.3	2,092	228.0
430	698	4,380	52.0	910	329.7	2,070	233.2
440	682	4,280	54.5	920	326.1	2,047	238.1
450	667	4,190	57.0	930	322.6	2,024	243.4
460	652	4,100	59.6	940	319.1	2,003	248.7
470	638	4,010	62.3	950	315.8	1,982	254.1
480	625	3,920	64.8	960	312.5	1,962	259.5
490	612	3,842	67.6	970	309.3	1,942	264.7
				980	306.1	1,922	270.4
500	600	3,766	70.4	990	303.0	1,902	275.9
510	588	3,692	73.3				
520	577	3,620	76.0	1,000	300.0	1,884	281.6
530	566	3,552	79.0	1,050	285.7	1,794	310.5
540	556	3,485	82.1	1,100	272.7	1,712	340.4
550	545	3,422	85.2	1,150	260.9	1,637	372.1
560	536	3,361	88.4	1,200	250.0	1,570	405
570	526	3,302	91.4	1,250	240.0	1,506	440
580	517	3,246	94.7	1,300	230.8	1,448	476
590	509	3,193	98.0	1,350	222.2	1,395	513
				1,400	214.4	1,346	552
600	500	3,140	101.4	1,450	206.9	1,298	592
610	492	3,088	104.7				
620	484	3,038	108.2	1,500	200.0	1,256	634
630	476	2,990	111.7	1,550	193.5	1,215	676

TABLE XV.—RELATION OF WAVE LENGTH (λ), FREQUENCY (f), $2\pi f = (\omega)$, AND $L \times C$.
(Concluded)

λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm. C in μf .	λ Wave length, meters	f Multiply values below by 1,000	ω Multiply values below by 1,000	LC L in cm. C in μf .
1,600	187.5	1,177	720	2,900	103.5	648	2,366
1,650	181.8	1,142	766	2,950	101.7	638	2,450
1,700	176.5	1,108	813				
1,750	171.4	1,076	862	3,000	100.0	628	2,533
1,800	166.7	1,046	912	3,500	85.7	538	3,448
1,850	162.2	1,017	963	4,000	75.0	471	4,500
1,900	157.9	990	1,016	4,500	66.7	418	5,700
1,950	153.8	965	1,071	5,000	60.0	377	7,040
				5,500	54.5	342.2	8,520
2,000	150.0	942	1,126	6,000	50.0	314.2	10,140
2,050	146.3	920	1,183	6,500	46.2	289.8	11,880
2,100	142.9	898	1,241	7,000	42.9	268.8	13,780
2,150	139.5	876	1,301	7,500	40.0	251.0	15,830
2,200	136.4	856	1,362				
2,250	133.3	838	1,425	8,000	37.50	235.2	18,010
2,300	130.4	819	1,489	8,500	35.29	221.4	20,340
2,350	127.7	801	1,555	9,000	33.33	209.2	22,800
2,400	125.0	784	1,622	9,500	31.58	198.2	25,410
2,450	122.5	768	1,690	10,000	30.00	188.4	28,160
				15,000	20.00	125.7	63,400
2,500	120.0	753	1,760	20,000	15.00	94.2	112,600
2,550	117.7	738	1,831	25,000	12.00	75.4	176,000
2,600	115.4	724	1,903	30,000	10.00	62.8	253,300
2,650	113.2	710	1,977	35,000	8.57	53.8	344,800
2,700	111.1	697	2,052				
2,750	109.1	684	2,129	40,000	7.50	47.1	450,000
2,800	107.1	672	2,207	45,000	6.67	41.8	570,000
2,850	105.3	660	2,287	50,000	6.00	37.7	704,000

Directions for Using Reactance and Resonant-Frequency Charts¹

(Charts appear on inside of back cover)

The charts appearing on the inside of the back cover are intended to assist in the solution of alternating-current problems and may be used to find:

1. The reactance of a given inductance at a given frequency.
2. The reactance of a given capacitance at a given frequency.
3. The resonant frequency of a given inductance and capacitance.

The charts cover all values of reactance, inductance, capacitance, and frequency encountered in practice, and automatically and properly locate the decimal point. For convenience in determining the approximate magnitude of frequency-wave-length conversion, a wave-length scale is indicated (chart A) in addition to the frequency scale.

In order that the user may determine the proper magnitude of the quantities involved to two or three significant figures with a minimum of trouble and without recourse to the use of instruments, chart A is divided into two parts. Chart A is the complete chart to be used for rough calculations whereas chart B, which is a single

¹ Courtesy of General Radio Company, Cambridge, Mass.

decade of chart *A* enlarged approximately seven times, is to be used where the significant figures are to be determined to two or three places. All scales are logarithmic, with the 1 and 5 coordinates indicated in chart *A*.

The proper use of the charts may best be illustrated by the following examples.

1. *Given the inductance and the frequency: find the reactance.*

Locate the given frequency on the horizontal frequency scale at the bottom of chart *A*, noting whether the upper or the lower set of numerals is used. Project vertically to the given value of inductance on the slant inductance scale, using the upper or lower set of numerals on this scale in accordance with those used on the frequency scale. From the intersection of the frequency and inductance values thus determined, project to the left and read the value of inductive reactance from the reactance scale. A more accurate numerical value of reactance can be found from chart *B* by following the same procedure as outlined above for chart *A*.

Example: An inductance of 0.5 henry (upper inductance scale) at a frequency of 712 kc. (upper frequency scale) has a reactance of roughly 2,000,000 ohms, as indicated in chart *A* by the cross in the upper right corner. From chart *B*, the reactance of a 0.5-henry coil at 712 kc. is found to be, more accurately, 2,240,000 ohms. Similarly, by using the lower set of numerals on the inductance and frequency scales an inductance of 0.5 microhenry at 712,000 Mc. has a reactance of 2,240,000 ohms.

2. *Given the capacitance and the frequency: find the reactance.*

On chart *A*, locate the given frequency on the horizontal frequency scale, noting whether the upper or lower set of numerals is used. Project vertically to the given value of capacitance on the slant capacitance scale, using the upper or lower scale, depending upon whether the upper or lower frequency scale, respectively, was used. From the intersection of the frequency and the capacitance values thus determined, project to the left and read the approximate value of capacitive reactance from the reactance scale. A more accurate numerical value of reactance can be found from chart *B*, by following the same procedure as outlined above for chart *A*.

Example: A capacitance of 0.1 $\mu\text{mf.}$ at 712 kc. (upper set of values on both scales) has a capacitive reactance of roughly 2,000,000 ohms. From chart *B*, the reactance, accurate to three significant figures, is found to be 2,240,000 ohms.

3. *Given the inductance and the capacitance: find the resonant frequency.*

Locate both the given values of inductance and capacitance (on the slant coordinate axes) on chart *A* noting whether the upper or lower set of values is used. Project down to the frequency scale, and read the upper frequency scale, if upper scale values were used for inductance and capacitance, or the lower frequency scale, if lower scale values were used for inductance and capacitance. Chart *B* is used in the same way to give the numerical solution to several significant figures.

Example: A capacitance of 0.1 $\mu\text{mf.}$ in combination with an inductance of 0.5 henry will have a resonant frequency of roughly 700 kc. as determined from chart *A*. The first three significant figures are found, from chart *B*, to be 712, making the resonant frequency 712 kc. By using the lower scale values, the resonant frequency of 0.1 $\mu\text{mf.}$ in combination with 1 $\mu\text{h.}$ is also found to be 712 kc.

Caution: Always obtain an approximate value from chart *A* before obtaining a more accurate value from chart *B*. Failure to observe this precaution may lead to erroneous results because the manner in which the chart necessarily represents the mathematical relations between the four variables involved, requires two logarithmic decades for inductance and capacitance, for every single decade of frequency and reactance. Unless the correct decade for the inductance and the capacitance is chosen, the calculated values of reactance and frequency will be in error by a factor of approximately 3.

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APPENDIX B

USEFUL OPERATING INFORMATION

LIST OF ABBREVIATIONS TO BE USED IN RADIO COMMUNICATIONS

I. Q. CODE

ABBREVIATIONS TO BE USED IN ALL SERVICES^{1,2}

Abbreviation	Question	Answer or statement
QRA	What is the name of your station?	The name of my station is
QRB	At what approximate distance are you from my station?	The approximate distance between our stations is nautical miles (or kilometers).
QRC	By what private operating enterprise (or government administration) are the accounts for charges of your station settled?	The accounts for charges of my station are settled by the private operating enterprise (or by the government administration of).
QRD	Where are you going and where do you come from?	I am going to and I come from
QRG	Will you tell me what my exact frequency (wave length) is in kilocycles (or meters)?	Your exact frequency (wave length) is kilocycles (or meters).
QRH	Does my frequency (wave length) vary?	Your frequency (wave length) varies
QRI	Is the tone of my transmission regular?	The tone of your transmission varies.
QRJ	Are you receiving me badly? Are my signals weak?	I cannot receive you. Your signals are too weak.
QRK	Are you receiving me well? Are my signals good?	I am receiving you well. Your signals are good.
QRL	Are you busy?	I am busy (or I am busy with). Please do not interfere.
QRM	Are you being interfered with?	I am being interfered with.
QRN	Are you troubled by static?	I am troubled by static.
QRO	Must I increase power?	Increase power.
QRP	Must I decrease power?	Decrease power.
QRQ	Must I transmit faster?	Transmit faster (. . . . words per minute).
QRS	Must I transmit more slowly?	Transmit more slowly (. . . . words per minute).
QRT	Must I stop the transmission?	Stop the transmission.
QRU	Have you anything for me?	I have nothing for you.
QRV	Are you ready?	I am ready.
QRW	Must I advise that you are calling him on kilocycles (or meters)?	Please advise that I am calling him on kilocycles (or meters).
QRX	Must I wait? When will you call me again?	Wait (or wait until I have finished communicating with). I shall call you again at o'clock (or immediately).
QRY	Which is my turn?	Your turn is number (or according to any other indication).
QRZ	By whom am I being called?	You are being called by
QSA	What is the strength of my signals (1 to 5)?	The strength of your signals is (1 to 5).
QSB	Does the strength of my signals vary?	The strength of your signals varies.

¹ Abbreviations take the form of questions when they are followed by a question mark.

² The series of signals QA, QB, QC, QD, QF, QG are reserved for the special code of the aeronautical service.

ABBREVIATIONS TO BE USED IN ALL SERVICES.^{1,2}—(Continued)

Abbreviation	Question	Answer or statement
QSD	Is my keying correct; are my signals distinct?	Your keying is incorrect; your signals are bad.
QSG	Must I transmit telegrams (or one telegram) at a time?	Transmit telegrams (or one telegram) at a time.
QSJ	What is the charge to be collected per word to including your internal telegraph charge?	The charge to be collected per word to is francs, including my internal telegraph charge.
QSK	Must I continue the transmission of all my traffic; I can hear you between my signals?	Continue the transmission of all your traffic; I shall interrupt you if necessary.
QSL	Can you acknowledge receipt?	I am acknowledging receipt.
QSM	Must I repeat the last telegram which I transmitted to you?	Repeat the last telegram which you transmitted to me.
QSO	Can you communicate with directly (or through)?	I can communicate with directly (or through).
QSP	Will you relay to free of charge?	I will relay to free of charge.
QSR	Has the distress call received from been attended to?	The distress call received from has been attended to by
QSU	Must I transmit (or answer) on kilocycles (or meters) and/or on waves of type A1, A2, A3, or B?	Transmit (or answer) on kilocycles (or meters) and/or on waves of type A1, A2, A3, or B.
QSV	Must I transmit a series of V's?	Transmit a series of V's.
QSW	Do you wish to transmit on kilocycles (or meters), and/or on waves of type A1, A2, A3, or B?	I am going to transmit (or I shall transmit) on kilocycles (or meters), and/or on waves of type A1, A2, A3, or B.
QSX	Will you listen to (call signal) on kilocycles (or meters)?	I am listening to (call signal) on kilocycles (or meters).
QSY	Must I shift to transmission on kilocycles (or meters), without changing the type of wave? or Must I shift to transmission on another wave?	Shift to transmission on kilocycles (or meters) without changing the type of wave, or Shift to transmission on another wave.
QSZ	Must I transmit each word or group twice?	Transmit each word or group twice.
QTA	Must I cancel telegram No. . . . as if it had not been transmitted?	Cancel telegram No. . . . as if it had not been transmitted.
QTB	Do you agree with my word count?	I do not agree with your word count; I shall repeat the first letter of each word and the first figure of each number.
QTC	How many telegrams have you to transmit?	I have telegrams for you (or for).
QTE	What is my true bearing in relation to you? or What is my true bearing in relation to (call signal)? or What is the true bearing of (call signal) in relation to (call signal)?	Your true bearing in relation to me is degrees or Your true bearing in relation to (call signal) is degrees at (time) or The true bearing of (call signal) in relation to (call signal) is degrees at (time).
QTF	Will you give me the position of my station on the basis of bearings taken by the radio direction-finding stations which you control?	The position of your station on the basis of bearings taken by the radio direction-finding stations which I control is latitude, longitude.
QTG	Will you transmit your call signal during 50 sec. ending with a 10-sec. dash, on kilocycles (or meters) so that I may take your radio direction-finding bearings?	I will transmit my call signal during 50 sec., ending with a 10-sec. dash, on kilocycles (or meters) so that you may take my radio direction-finding bearings.
QTH	What is your position in latitude and in longitude (or according to any other indication)?	My position is latitude, longitude (or according to any other indication.)
QTI	What is your true course?	My true course is degrees.
Q TJ	What is your speed?	My speed is knots (or kilometers) per hour.

ABBREVIATIONS TO BE USED IN ALL SERVICES^{1,2}—(Concluded)

Abbreviation	Question	Answer or statement
QTM	Transmit radio signals and submarine sound signals to enable me to determine my bearing and my distance.	I am transmitting radio signals and submarine sound signals to enable you to determine your bearing and your distance.
QTO	Have you left dock (or port)?	I have left dock (or port).
QTP	Are you going to enter dock (or port)?	I am going to enter dock (or port).
QTQ	Can you communicate with my station by the International Code of Signals?	I am going to communicate with your station by the International Code of Signals.
QTR	What is the exact time?	The exact time is
QTU	What are the hours during which your station is open?	My station is open from to
QUA	Have you any news from (call signal of the mobile station)?	This is the news from (call signal of the mobile station).
QUB	Can you give me, in the following order, information concerning: visibility, height of clouds, ground wind at (place of observation)?	This is the information requested:
QUC	What is the last message you received from (call signal of the mobile station)?	The last message I received from (call signal of the mobile station) is
QUD	Have you received the urgent signal transmitted by (call signal of the mobile station)?	I have received the urgent signal transmitted by (call signal of the mobile station) at (time).
QUF	Have you received the distress signal sent by (call signal of the mobile station)?	I have received the distress signal sent by (call signal of the mobile station) at (time).
QUG	Will you be forced to come down on water (or on land)?	I am forced to come down on water (or on land) at (place).
QUH	Will you give me the present barometric pressure at sea level?	The present barometric pressure at sea level is (units).
QUJ	Will you please indicate the proper course to steer towards you, with no wind?	The proper course to steer towards me, with no wind, is degrees at (time).

2. MISCELLANEOUS ABBREVIATIONS

Abbreviation	Meaning
C	Yes.
N	No.
P	Announcing private telegram in the mobile service (to be used as a prefix).
W	Word or words.
AA	All after (to be used after a question mark to request a repetition).
AB	All before (to be used after a question mark to request a repetition).
AL	All that has just been transmitted (to be used after a question mark to request a repetition).
BN	All between (to be used after a question mark to request a repetition).
BQ	Answer to RQ.
CL	I am closing my station.
CS	Call signal (to be used in requesting that call signal be given or repeated).
DB	I cannot give you a bearing, you are not in the calibrated sector of this station.
DC	The minimum of your signal is suitable for the bearing.
DF	Your bearing at (time) was degrees, in the doubtful sector of this station, with a possible error of two degrees.
EG	Please advise me if you find an error in the bearing given.
DI	Doubtful bearing due to the bad quality of your signal.
DJ	Doubtful bearing due to interference.

MISCELLANEOUS ABBREVIATIONS—(Concluded)

Abbreviation	Meaning
DL	Your bearing at (time) was degrees, in the uncertain sector of this station.
DO	Doubtful bearing. Request another bearing later, or at (time).
DP	Beyond 50 miles, possible error of bearing can attain two degrees.
DS	Adjust your transmitter; your minimum signal is too broad.
DT	I cannot give you a bearing; your minimum signal is too broad.
DY	This is a two-way station; what is your approximate direction, in degrees, in relation to this station?
DZ	Your bearing is reciprocal (to be used only by the control station of a group of radio direction-finding stations when addressing other stations of the same group).
ER	Here (to be used before the name of the mobile station in the transmission of routing indications).
GA	Resume transmission (to be used more especially in the fixed service).
JM	If I may transmit, make a series of dashes. To stop my transmission, make a series of dots [not to be used on 500 kc. (600 m.)].
MN	Minute or minutes (to be used to indicate the duration of the waiting period).
NW	I am resuming transmission (to be used more especially in the fixed service).
OK	We agree.
RQ	Announcing a request.
SA	Announcing the name of an aircraft station (to be used in transmitting transit data).
SF	Announcing the name of an aeronautical station.
SN	Announcing the name of a coast station.
SS	Announcing the name of a ship station (to be used in transmitting transit data).
TR	To announce sending of indications concerning a mobile station.
UA	Do we agree?
WA	Word after (to be used after a question mark to request a repetition)
WB	Word before (to be used after a question mark to request a repetition).
XS	Static.
YS	See your service notice.
ABV	Repeat (or I repeat) the figures in abbreviated form.
ADR	Address (to be used after a question mark to request a repetition).
CFM	Confirm (or I confirm).
COL	Collate (or I collate).
ITP	The punctuation counts.
MSG	Announcing a telegram concerning the service on board (to be used as a prefix).
NIL	I have nothing to transmit to you (to be used after an abbreviation of code Q to show that the answer to the question asked is in the negative).
PBL	Preamble (to be used after a question mark to request a repetition)
REF	Reference to (or Refer to).
RPT	Repeat (or I repeat) (to be used in requesting or giving repetition of all or part of traffic, the abbreviation to be followed by the corresponding indications).
SIG	Signature (to be used after a question mark to request a repetition).
SVC	Announcing a service telegram concerning private traffic (to be used as a prefix).
TFC	Traffic.
TXT	Text (to be used after a question mark to request a repetition).

SCALE USED TO EXPRESS STRENGTH OF SIGNALS

(To be used following the signal QSA)

- 1 = scarcely perceptible; unreadable.
- 2 = weak; readable now and then.
- 3 = fairly good; readable, but with difficulty
- 4 = good; readable.
- 5 = very good; perfectly readable.

RADIO WEATHER TRANSMISSION CODE

The standard teletypewriter symbols and their code equivalents as used by the radio-telegraph stations are shown below. The radio characters representing symbols will be transmitted without space, for example: ---...--; and not --- ...--.

TELETYPEWRITER SYMBOL	RADIO-TELEGRAPH EQUIVALENT
+	(Heavy, dense, severe) •-•-• (AR) (cross)
/	(Slant, for separation) -••-• (DN)
○	(Clear) - - - • (German O)
⊕	(Scattered clouds) - - - • - - - (Exclamation point)
⊕	(Overcast) - - - - - • • • (OX)
⊕	(Broken clouds) - • • • - • • • - (BX)
-	(Light, thin mild) - • • • • • - (DX)

INTERNATIONAL MORSE CODE

A	• ■■	S	• • •
B	■■ • • •	T	■■■
C	■■ • ■■ •	U	• • ■■
D	■■ • •	V	• • • ■■
E	•	W	• ■■ ■■
F	• • ■■ •	X	■■ • • ■■
G	■■ ■■ •	Y	■■ • ■■ ■■
H	• • • •	Z	■■ ■■ • •
I	• •	1	• ■■ ■■ ■■ ■■
J	• ■■ ■■ ■■	2	• • ■■ ■■ ■■
K	■■ • ■■	3	• • • ■■ ■■
L	• ■■ • •	4	• • • • ■■
M	■■ ■■	5	• • • • •
N	■■ •	6	■■ • • • •
O	■■ ■■ ■■	7	■■ ■■ • • •
P	• ■■ ■■ •	8	■■ ■■ ■■ • •
Q	■■ ■■ • ■■	9	■■ ■■ ■■ ■■ •
R	• ■■ •	0	■■ ■■ ■■ ■■ ■■

PUNCTUATION AND SPECIAL SIGNALS

.	• • • • •	/	■■ • • ■■ •
=	■■ • • • ■■	?	• • ■■ ■■ • •
Comma (,) -----	• ■■ • ■■ • ■■		
Parenthesis [()] -----	■■ • ■■ ■■ • ■■		
From (de) -----	■■ • • •		
Go Ahead (k) -----	■■ • ■■		
Received O. K. (R) -----	• ■■ •		
End of Message (cross x) -----	• ■■ • ■■ •		
Transmission Finished (sk) -----	• • • ■■ • ■■		

INDEX

A

- AVC* (see Automatic volume control)
- Abbreviations, code, 739-742
- Absolute zero temperature, 12
- Absorption coefficient, 310-311
- Accepter circuits, 93, 242, 380
- Acoustics, 310-314
- Adcock antenna, 537, 564
- Admittance, 97
- Aerial (see Antenna)
- Aircraft antennas, 297-298
- Aircraft bonding, 507-510
- Aircraft direction finders, 551-552, 555, 559, 570-574
- Aircraft receivers, 475-483, 499-506, 570-574
- Aircraft rectifier unit, 460-461
- Aircraft shielding, 507-510
- Aircraft signal indicators, 567-574
- Aircraft transmitters, 456-460, 461-471, 562-570
- All-wave receiver, 475-483
- Alternating current, circuit analysis, 58-59, 83-115
- Alternating-current generators, 58-60, 617-620
- Alternating-current meters, 663-669
- Alternating-current motors, 636-638
- Alternator, simple, 58-60
 three-phase, 111-115
 types of, 617-620
- Ammeters, 661-663, 665-668
 alternating current, 665
 direct current, 661
 hot wire, 665-668
 thermal, 665-668
- Ampere, 10
- Ampere-hour capacity, 33, 683
- Ampere-hour meter, 668, 707-710
- Ampere turns, 52
- Amplification factor, 122-123
- Amplifier calculations, 315-321
- Amplifiers, 127-135, 214-233
 audio-frequency, 216-232, 357, 403-407
- Amplifiers, buffer, 169-170
 Class A, 131-133
 Class B, 133-134, 169-173, 188-191
 Class C, 134, 169-172, 180-181
 design of, in transmitters, 169-175, 179-193
 distortion in (see Distortion)
 efficiency of, 132, 134
 frequency-doubling, 172-175
 hum in, 124-126, 374-375
 impedance coupled, 220-221, 365
 linear, 134, 172, 183, 189, 212-214
 output impedance adjustment of, 184-185, 191-194
 power (see Power amplifiers)
 push-pull, 134, 138, 227-232
 radio-frequency, 127-135, 158-191
 resistance coupled, 218-220, 364-365
 speech, 327, 333, 348, 357-383, 403-407
 transformer coupled, 216-218, 365-367
- Amplitude, 83-85
 distortion, 359-362
 modulation, 175-178
- Antenna, Adcock, 537, 564
 aircraft, 297-298, 506-507
 broadcast, 294-296
 calculations, 267-268, 283-285, 288-289, 306-308
 coupling units, 194-195, 287-292, 415
 current-feed, 284-287
 dipole, 261-262
 directional, 537, 562-567
 excitation, 282-287
 field patterns, 280, 524, 534, 562, 566, 568, 570-571
 Franklin, 281-282
 half-wave doublet, 261-262
 Hertz, 261-262
 loop, 301, 523-562
 marine, 296
 measurements, 267-271, 302-308
 quarter-wave, 362-364
 RCA broadside, 279-281
 radiation, 265-282
 reactance curves, 264

- Antenna, reactance measurements, 306-308
 short-wave, 277-279, 292-293
 T-L, 562-567
 transmission lines, 282-291, 562-567
 ultra-high frequency, 293-294
 voltage-feed, 284-287
 Archimedes' principle, 681
 Armatures, 617-628
 testing of, 652
 Array, antenna, 537, 562-567
 Astatic pick up, 340-342
 Atoms, 3-5
 Attenuation networks, 349-357, 384-390
 Audio-frequency, oscillator, 425, 436, 440
 (See also Amplifiers)
 Audio spectrum chart, 313
 Automatic relays, 397-399, 653-656
 Automatic starters, 638-647
 Automatic temperature control, 167-169
 Automatic volume control, 235-241
 Automobile receivers, 511-518
- B
- B-H curves, 53-55
 Balance circuits, 134-139, 227-232
 Band-pass filter, 246-249
 Band width, 177-178, 242-244, 247-249
 Barkhausen-Kurz oscillator, 204-205
 Batteries (see Cells)
 Battery charging, 437, 683-695, 700
 Battery panels, 437, 702-707, 711
 Bearings, radio (see Direction finders)
 Bellini-Tosi antenna, 535-537
 Biasing (see Grid-biasing)
 Bibliography, 737-738
 Bilateral characteristic, 523-527
 Bimorph element, 332
 Bonding, 509-510
 Bridge circuits, high frequency, 134-139
 rectifier, 586-587
 Broad tuning, 130, 241-245
 Broadcast antennas, 294-296, 562-567
 Broadcast equipment, 345-384, 321-344,
 396-420, 687-697
 emergency battery, 687-697
 power, 369-370, 373-376, 687-697
 studio, 321-344, 403-407
 Broadcast studio (see Studio)
 Broadcast transmitters, 391-420
 circuit theory of, 155-206
 tuning of, 399-403
 Buffer amplifier, 169-170
- C
- CW (see Continuous wave)
 Capacitance, 76-83
 Capacitive coupling, 136-138, 149, 161,
 187, 195, 199, 219, 248
 Capacitive reactance, 90-91
 Capacity, of a cell, 33
 electrostatic, 76-83
 neutralization of, in tubes, 135-139,
 399-403
 Carbon microphone, 36-38, 321-337
 Cardioid characteristic, 534
 Cathode-ray oscillograph, 412-414
 Cell connections, 33-34
 Cells, 671-712
 capacity of, 33, 683-684, 701
 care of, 695-697
 charging of, 680, 683-695, 700-707,
 710-712
 chemical action in, 29-31, 680, 699-700
 dry, 35-36
 Edison, 697-701
 electrolyte in, 675-677, 697-700
 Exide, 672-697
 lead-acid, 672-697
 methods of connecting, 33-34
 Planté type, 672-674
 primary, 29-30, 35-36
 secondary, 671-712
 temperatures of, operating, 682, 690-
 691
 Characteristic impedance, 283
 Characteristics, class A amplifier, 132-133
 , class B amplifier, 133-134
 class C amplifier, 134
 microphone, 343
 vacuum tube, 121-124
 Charged bodies, 2-6
 Charging, rates of, 690, 700-701
 requirements of, 686-691
 Charging panels, broadcasting, 687-697
 marine, 437, 687, 702-707
 Charging storage cells, 684-696, 700-707,
 710-712
 Charts, inductance, 74
 radio-frequency spectrum, 158
 vacuum tube, 714-731
 Chemical action, 29-31
 in primary cells, 29-32
 in secondary cells, 680, 699-700
 Chloride accumulator (see Storage bat-
 teries)

- Choke coils, low-frequency, 606-607
 radio-frequency, 144-147
- Chopper, 424
- Class A amplifier, 131-133
- Class B amplifier, 132-134, 169-173, 188-191
- Class C amplifier, 134, 169-172, 180-181
- Classification, radio spectrum, 155-156
- Coast Guard, direction finders used by, 547-557
 frequency meter used by, 519-522
 receivers used by, 487-489
 transmitters used by, 443-446
- Coastal stations, 453
- Code, abbreviations, 739-743
 chart, 744
- Coefficient, coupling, 76
- Coercive force, 56
- Colpitts oscillator, 149-150
- Common impedance, 144-145, 232
- Commutators, 621-628
- Compass, radio (*see* Direction finder)
- Complex, circuits, 21-26
 quantities, 104-109
- Condenser, 76-83, 604-606
 microphone, 326-329
- Conductance, 26, 98-99
 mutual, 124
- Conduction, electrical, 2, 6-9
- Constant current modulation, 181-184
- Continuous waves, production of, 147-154
 receivers of, 250-251, 255, 257-260
 transmitters of, 421-474
- Control-room equipment, 345-384, 403-407
- Conventional current flow, 31
- Conversion table, frequency-wave length, 732-735
- Cooling system, vacuum tube, 419-420
- Copper oxide rectifiers, 596-599
- Coulomb, 10-11
 law of, 6
- Counter e.m.f., 67, 633
- Counterpoise, 298-299
- Coupling, in audio-frequency circuits, 216-221
 capacitive, 195, 199, 248
 magnetic, 69-70, 75-76
 units of, 287-292, 415
 in radio-frequency circuits, 127-131, 169-175, 179-186, 241-250, 287-291
 (*See also* Antenna coupling units)
- Coupling coefficient, 76
- Cross modulation (cross talking), 209
- Crystal-controlled oscillator, 160, 164-169
- Crystal-controlled transmitter, 160-161
 frequency monitor of, 407-410
- Crystal filter, 258-260
- Crystal microphone, 332-336
- Crystal resonator, 164-167
 temperature control of, 167-169
- Current, alternating, 83-85
 direct, 7-9, 31-32
 electric, 7-9, 83-85
 indicating instruments for, 661-663, 665-668, 707-710
- Current-feed antenna, 284-287
- Current flow, 31-32
- Current-square ammeter, 666
- Cycle, 60
- D
- Decibel, 315-321
 levels, 318
 ratio table, 316, 386
- Delta transformer connections, 111, 589
- Detection, 209-214
 diode, 233-235
 double, 253-260
 grid bias, 212-214
 grid leak, 211-212
 linear, 212-214
 plate, 210-211
 power, 212-214
- Diamagnetic, 43
- Diode detector, 233-235
- Dipoles (*see* Antennas)
- Direct-current generators, 620-631
- Direct-current meters, 659-663, 669-670, 707-710
- Direct-current motors, 631-636
- Direction of current flow, 31-32
- Direction finder, 523-574
 Adcock system, 537, 564
 aircraft, 551-554, 570-574
 antenna effect, 528-532
 Bellini-Tosi, 535-537
 bilateral characteristic, 523-527
 calibration, 553-555
 cardioid characteristic, 533-534
 deviation, 528-532
 directional characteristics, 523-528
 errors, 528

- Direction finder, fundamental, 301, 523,
 541-545
 marine, 523-551, 553-560
 night effect, 527, 560-562
 plotting a position with, 558-559
 practical operation, 555-559
 sense characteristic, 533, 574
 signal indication, 546, 567-570
 troubles, 560
 unilateral characteristic, 533
 Directional antennas, 281-282, 523-528,
 537, 562-567
 Distortion, amplifier, 193, 196-200, 221-
 223, 309, 359-362
 Doublet antenna, 261-262, 277
 tuning a, 278
 (See also Antenna)
 Dry cells, 35-36
 Dynamic microphone, 329-330
 Dynamo (see Generators)
 Dynamotor, 647-648
 Dynatron oscillator, 153-154
- E
- E*-layer, 273
 Echo signals, 275
 Eddy currents, 62
 Efficiency, of Class A amplifier, 132
 of Class B amplifier, 134
 of Class C amplifier, 134
 Electric charges, 1-6
 Electric circuit, 15-26
 Electric current, 7-11
 Electric field, 5, 76-77, 265
 Electric forces, 1-6
 Electric meters, 195, 659-670, 707-710
 Electric potential, 5-6
 Electric power, 26-28, 109-115
 Electrical-transcription apparatus, 338-
 344
 Electrolytic condensers, 604-606
 Electromagnetic fields, 66-70, 265
 Electromagnetic induction, 66-70
 Electromagnetism, 38-57
 Electromotive force (e.m.f.), 9-10, 29-31,
 38, 49-51, 60-70
 effective, 87
 maximum, 88
 Electron-coupled oscillators, 151, 152, 204
 Electron flow, 6-10
 in conductors, 6-10
 in gases, 117, 118
 Electron flow, in liquids, 30-32
 Electron mass, 3
 Electrons, free, 6-7
 in magnetic materials, 42-43
 velocity of, 7-8
 Electroscope, 2
 Electrostatic capacity, 76-83
 Electrostatic charges, 1-6
 Electrostatic field, 2-5, 76-77, 265
 Electrostatic forces, 1-6
 Electrostatic shielding, 545, 578
 Electrostatic units, 3
 Emergency marine batteries, 686-687
 Emergency power supply, 702-705
 (See also Storage batteries)
 Emergency transmitter, 426-428
 Energy, 26-27
 Equalizers, 380-383
 Equipment, broadcasting, 345-384, 321-
 344, 391-420
 control room, 345-384
 studio, 321-344
 testing, 467, 651-653
 Ergs, 27
 Ether, 265
- F
- F*-layer, 273
 Farad, 78
 Ferromagnetic, 43
 Field, generator, excitation, 628
 Field strength, measurements of, 267-
 271
 Filament, 116-118
 Filter, 245-250
 band-pass, 246-247
 chokes for, 606-607
 design of, 599-614
 grid, 232-233
 high-pass, 246
 low-pass, 246, 601-602
 plate, 144-147, 232-233
 power-supply, 599-609
 ripple, 599-608, 657-659
 scratch, 342-343
 Filter condensers, 604-606
 Flux, magnetic, 52-53
 Flywheel effect, 67-68, 133-134
 Force, 1-6, 26
 Franklin antenna, 281-282
 Free electrons, 6-7
 Frequency, 60, 65, 85-87, 100, 285

- Frequency control, 167-169, 395-396, 407-410
- Frequency doublers (multipliers), 172-175
- Frequency meter, 519-522, 668-669
- Frequency monitor, 194, 407-410
- Frequency runs, 382-383
- G
- Gain calculations, 315-321
- Gain control, 368
(*See also* Volume controls)
- Galvanometer, 50, 659-661
- Gauss, 53
- Generation of e.m.fs., chemical, 29-31
induction of, 38, 49-51, 60-70
- Generators, alternating current, 58-60, 617-620
direct current, 620-631
elementary principles of, 50-51, 58-60
regulation of, 630-631
voltage control on, 629-630
- Gilbert, 53
- Gill-Morrell oscillations, 205
- Grid-biasing, 120-121, 132-134
methods of, 142-144, 166, 211-212, 370-372
- Grid excitation, in transmitters, 186-191
- Grid filtration, 145, 232-233
- Grid modulation, 179-180
- Ground waves, 266
- Gutton-Touly oscillator, 203
- H
- Half wave antennas, 261-262, 294-296
- Hand rules, 49, 51
- Harmonic producers, 172-175
- Harmonics, 85, 196-200, 359-362
suppression of, 193, 198-200
- Hartley oscillator, 149
- Hazeltine balance (neutralization), 135-137
- Heising modulation, 181-184
- Helium atom, 4
- Hertz antenna, 261-262
- Heterodyne, 253-260
super-, 253-260
- Heterodyne frequency meter, 519-522
- High fidelity, 322
use of side bands, 177-178
- High-frequency telegraph transmitters, 428-434, 469-471
- High-frequency telephone transmitters, 456-460, 461-469
- High-frequency telephone-telegraph transmitters, 443-446
- High Q, 163
- Hot-wire ammeter, 667-668
- Hum in amplifiers, 124-126, 374-375
- Hydrogen atom, 4
- Hydrometer, 681-682
- Hysteresis, 56-57
- I
- ICW*, production of, 423-426
- IR* drop, 15
- Image frequency, 257, 259
- Impedance, 89-92
common, 232
coupling, 220-221, 365
load adjustments, 184-185, 191-194
parallel-resonant-circuit, 97, 184-185, 244
series-resonant-circuit, 93-95, 244
- Inductance, 70-76, 88-90
calculations, 72-76
- Induction, 49-51, 66-70
mutual, 66-70, 75-76
self, 66-70
- Inductive reactance, 88-89
- Inductively coupled circuits, 69-70, 129-131, 169-175, 179-186
- Inductor alternator, 619
- Inertia (*see* Flywheel effect)
- Insertion loss, 354
- Instruments, indicating, 659-670, 707-710
- Insulators, 2
- Internal drop, 15
- Internal resistance, 30, 34-35
- Inverse-peak voltage, 583-584
- Inverse-square law, 6
- Inverted L antenna, 262, 263, 296
- Ionization, 30-31, 582
- Ions, 30-31
- Isolation circuits, 144-147, 232-233, 247-250
- J
- J* operator, 100-109
- Joule, 27, 81
- Junction rectifiers, 596-599

K

Kelvin temperature, 12
 Kennelly-Heaviside layer, 272-276
 effects of, 560-561
 Keying methods, 422-423, 426
 Kilovolt-ampere, 110
 Kinetic energy, 68, 133-134
 Kirchoff's law, 22-26

L

Lapel microphone, 336
 Laws, inverse-square, 6
 Kirchoff's, 22-26
 Lenz's, 49-50, 68-69
 Ohm's, 13-15, 53, 92-93
 Lead-acid cell, 672-697
 Left-hand rules, 48-49, 51
 Lecher wires, 285-287
 Lenz's law, 49-50, 68-69
 Line reflections, 282-283
 Linear amplifier, 134, 172, 183, 189, 212-214
 Linear detector, 212-214
 Load-line curves, 192
 Load matching, 28-29, 122, 184-186, 191-194, 283
 Load voltage, 16
 Local action, 32
 Logarithms, 314-321, 756-758
 Long-wave attachment, 496-498
 Loop antenna, 301
 (See also Direction finder)
 Low-pass filters, 246, 601-602

M

M.O.P.A. transmitters, 158-160, 434-441
 Magmotor, 515
 Magnesium-copper-sulphide rectifier, 599
 Magnetic circuit, 51-57
 Magnetic coupling, 69-70, 75-76
 Magnetic fields, 38-52, 66-70, 265
 Magnetic materials, 52-56
 Magnetic pick-ups, 338-340
 Magnetic poles, 39
 Magnetic shielding, 139-142
 Magnetism, 38-57
 Magnetomotive force, 52
 Magnetostriction, 43, 45-46
 Magneto-oscillator, 163-164

Marconi antenna, 262-264
 Marine charging panels, 437, 702-707
 Marine direction finders, 523-551, 553-560
 Marine receivers, 483-499
 Marine transmitters, 396-410, 421-453
 Matter, 3
 Mechanical oscillator, 163-164
 Medium-frequency telegraph transmitters, 434-441, 469-471
 Mercury-arc rectifier, 593-596
 Mercury-vapor rectifier, 581-584
 Mesney oscillator, 203
 Meters, electrical, 195, 659-670, 707-710
 Microphone, 36-38, 321-337
 amplifiers of, 327, 333, 404
 carbon, 36-38, 323-326
 condenser, 326-329
 crystal, 332-336
 curves of, 343
 dynamic, 329-330
 lapel, 336
 placement of, 311-314
 ribbon, 330-331
 Microphone reflectors, 336-337
 Microphones, 36-38, 321-337
 Mixers, 349-357
 Modulating amplifier, 179-186
 Modulating methods, 179-184
 Modulation, 175-178
 constant current system of, 181-183
 grid bias, 179-181
 Heising system of, 181-183
 high level, 179
 low level, 179
 Modulation factor, 178
 Modulation percentage, 176-177, 412
 Monitor amplifiers, 376-377
 audio-frequency, 194
 radio-frequency, 407-410
 Motor starters, 638-647
 Motors, alternating current, 636-638
 armature, 620-628
 compound wound, 633-634
 differential wound, 635
 direct current, 631-636
 hand rule for, 632
 induction, 636-638
 interpole, 635
 maintenance of, 648-651
 operation of, 645-647
 principles of, 631-638
 series type, 634

- Motors, shunt type, 634
 starters, 638-647
 testing of, 651-653
 torque in, 633
- Musical harmonics, 360-361
 spectrum chart, 313
- Mutual conductance, 124
- Mutual inductance, 66-71
- Mutual inductance measurements, 75-76
- N
- Nagaoka, formula of, 72
 inductance chart, 74
 table of, 73
- Negative resistance, 251
- Neutralizing transmitter, 138-139, 399-403
 (*See also* Balance circuits)
- Neutron, 4
- Nickel-iron-alkaline cell, 697-701
- Night effect, 527, 560-562
- Noise suppression, 239-241
 in amplifiers, 351-353
 antenna system, 300
 in automobile receivers, 515-517
 (*See also* Shielding and bonding)
 generator (*see* Ripple filters)
 motor (*see* Ripple filters)
- Non-magnetic materials, 52
- Nuclear electrons, 4
- O
- Oersted, 52
- Ohmmeters, 669-670
- Ohm's law, 13-15, 53, 92-93
- Oil-damped pick-up, 339-340
- Orbital electrons, 4
- Oscillators, 147-154, 157-169
 ultra-high-frequency, 201-206, 474
- Oscillograph, cathode ray, 412-414
- Overload circuit breakers (*see* Relays)
- P
- Pads, 349-357, 384-390
- Pan-American Airways, 469
- Parabolic microphone reflector, 336-337
- Parallel circuit, 16-18
- Parallel-feed, audio-frequency amplifier, 366-367, 404-406
 oscillator, 149-150, 162
- Parallel-resonance, 95-100, 241-245, 380-381
- Parallel-series circuit, 18-20
- Paramagnetic, 43
- Parasitic oscillations, 200-201, 367
- Peak factor, 87
- Peak voltage, 87-88
- Peak voltmeter, 411
- Pentagrid converter, 482
- Pentode, 225-227
- Permalloy, 54
- Permeability, 55
- Phase angle, 83-85, 91-92
- Phase relations, alternating-current circuits, 88-96
- Phonograph pick-up, characteristics of, 343
 devices used, 338-344
- Pick-up heads, phonograph, 338-342
- Piezo-astatic pick-up, 340-342
- Piezo-electric oscillator, 164-167
- Plate filtering, 144-147, 232-233
- Plate modulation, 181-184
- Plate resistance, 124
- Polarization, 30, 276-277
- Police, receivers, 511-518
 transmitters, 396-399, 456-460
- Polyphase circuits, 111
- Positive resistance, 251
- Positron, 4-5
- Potential, 5-6
- Potential difference, 5-6
 (*See also* Electromotive-force)
- Power, 26-29, 64, 109-115
- Power amplifiers, 170-191, 223-232
 distortion in, 196-200
 grid excitation in, 186-191
 harmonics in, 193, 196-200
 load calculations, 184-186, 191-194, 226-227, 288-289
- Power detection, 212-214
- Power factor, 92
- Power rectifier, 575-616
 (*See also* Rectifiers)
- Power supply, 196, 369-370, 687-697
- Power transformers, 42-66, 111-114, 575-579
- Primary cell, 29-30, 35-36
- Protective devices, 653-656
 (*See also* Relays)
- Protons, 3-4
 mass of, 3
- Push-pull amplification, 227-232
- Push-pull balance, 138
- Push-pull oscillator, 150

Q

- Q, of coil, 163, 248, 250
- Q, circuit, 163
- Q signals, 739-741
- QAVC (*see* Noise suppression)
- Quartz plate filter, 258-260
- Quartz plate oscillator, 160, 164-169

R

- RCA, modulation indicator, 412-414
 - radiophone attachment, 441-443
 - receivers, 475-483, 489-499
 - transmitters, broadcasting, 414-420
 - marine, 421-443, 447, 471-474
 - ultra-high-frequency, 471-474
 - tube charts, 714-727
- R.M.C.A. (*see* RCA)
- R.m.s. (root-mean-square) value, 87-88
- Radians, 85
- Radiating systems, 277-287, 291-300
 - (*See also* Antennas)
- Radiation, field, 265-282
 - resistance measurements, 302-308
- Radio aids to navigation, 523-574
- Radio-beacon, 560-574
 - charts, 539-541
 - receivers, 503, 570-574
 - signal indication, 567-570
 - transmitters, 560-567
- Radio compass (*see* Direction finder)
- Radio direction finder (*see* Direction finder)
- Radio-frequency (*see* Amplifiers)
 - choke coils, 144-147
 - spectrum chart, 158
 - spectrum classification, 155-156
- Radio-frequency amplifier, receiving, 127-134
 - transmitting, 169-175, 179-181, 184-191
- Radio goniometer, 535-537
- Radio-range beacons, 560-574
 - (*See also* Radio-beacon)
- Radio shadows, 275-276
- Radiophone attachment, marine, 441-443
- Reactance, 88, 90
 - charts, use of, 735-736
 - leakage, 61
- Receiver troubles, 498-499
- Receivers, aircraft, 475-483, 499-506
 - all-wave, 475-483

- Receivers, automobile, 511-517
 - marine, 483-499
 - police, 511-518
 - super-heterodyne, 253-260
- Receiving antenna, 300-301
- Receiving circuit analysis, 127-134, 241-260
- Receiving tubes, 714-719, 728-729
- Recording, 337-338
- Rectification (*see* Detection; Rectifiers)
- Rectifiers, bridge, 586-587
 - copper-oxide, 596-599
 - filtering, 599-609
 - full-wave, 584, 587-588, 592
 - half-wave, 583, 587, 590-591
 - losses, 614
 - magnesium-copper-sulphide, 599
 - mercury arc, 593-596
 - mercury vapor, 581-584
 - signal (*see* Detection)
 - theory of, 575-616
 - three-phase, 588-592
 - tubes, 727
 - voltage dividers for, 608-616
 - voltage doubling, 587-588
 - voltage regulation in, 577-578, 608-616
- Reflected resistance, 185, 244
- Reflection layer, 271-276
- Regeneration, 250-252
 - super-, 252-253, 518
- Regeneration control, 488, 495
- Rejector circuits, 249-250
- Relays, overload, 397-399, 653-656
 - thermal, 397-398
 - time delay, 375
 - vacuum-contact, 655-656
- Reluctance, 52, 55
- Remanence, 55
- Remote control, 383-384
- Residual magnetism, 55-56
- Resistance, 11-28
 - antenna, measurements, 302-308
 - coupled amplifier, 218-220, 364-365
 - internal, 34-35
 - load, 28-29
 - negative, 251
 - plate, 124
 - reflected, 185, 244
 - table, 12
- Resonance, 99-100
 - parallel, 95-100, 241-244, 380-381
 - series, 93-95, 241-244, 380-381
- Retentivity, 56

- Reverberation, 310
 Rheostats, 619, 628, 634, 638-640
 Ribbon microphone, 330-331
 Rice balance (neutralization), 138
 Ripple filters, 599-608, 657-659
 Ripple frequency, 613-614
 Rochelle-salts crystal, 332-336
 Rotary converter, 648
 Rules, left-hand, 48-49, 51
- S
- Sabine formula, 310
 Saturation, magnetic, 55
 tube, 118
 Scratch filters, 342-343
 Screen-grid tube, 207-209, 225-227
 Secondary cells (*see* Storage cells)
 Selectivity, 130, 241-245
 Sense characteristic, direction finder, 533-534, 555-557, 574
 Series circuit, 15-16
 calculations, 18
 Series feed, 147-149
 Series-feed oscillator, 161
 Series-parallel circuit, 18-20
 calculations, 19-22
 Series-resonance, 93-95, 99-109, 241-245
 Shielding, 139-142, 507-510, 515-517, 545
 Ship antennas, 296
 Ship receivers, 483-499
 Ship transmitters, 421-453
 Short-wave, antenna design, 277-287, 292-294
 Short-wave propagation, 271-277
 Short-wave receivers, 253, 475-483, 487-493, 499-501, 518
 Short-wave transmitters, 428-434, 443-446, 456, 470-474
 Shot effect, 362-363
 Shunt law, 21-22
 Side bands, 177-178
 (*See also* Band width)
 Sine wave, 83-85
 Single-signal receiver, 258-260
 Sinusoidal wave, 85
 Skip distance, 273-275
 Skip-distance curves, 274
 Sky waves, 266, 271-276
 Socket connections, receiving tubes, 720
 Sound, 314
 Space charge, 117
 Specific gravity, 681-682
 Specific inductive capacity, 79
 Specific resistance, 12
 Speech amplifiers, 327, 333, 348, 357-383, 403-407
 Speech-input equipment, 321-383, 403-407
 Standing waves, 264, 282-283, 293
 Star-transformer connections, 589
 Starting boxes, 638-647
 Statcoulomb, 3
 Stator, 637
 Storage batteries, 671-712
 (*See also* Cells)
 Edison, 697-701
 lead plate, 672-697
 Stray fields, 351-353
 Studio, bridging circuits in, 368-369
 microphones in, 321-337
 power supply in, 369-370, 373-76, 687-697
 remote control in, 383-384
 Studio acoustics, 309-314
 Studio amplifiers, 357-359, 403-407
 Studio apparatus, 309-344
 Sulphation, 693-695
 Superheterodyne, 253-260
 Superregeneration, 252, 518
 Surge impedance, 283
 Susceptance, 97
 Synchronous condenser, 636
- T
- T antenna, 262-264, 296
 T-L antenna (*see* Antenna, transmission-line)
 Telephone receiver, 36
 Temperature control, 167-169
 Testing, 649-653
 Thermal-relay, 397-398
 Thermocouple ammeter, 665-667
 Time constant, 237-239
 Time delay relay, 375
 Tone controls, 342, 349
 Torque, 633
 Transformer coupling (receiving), 216-218, 365-367
 Transformers, 60-66, 111-114, 575-579, 587-593
 connections of, 577, 584, 587, 588-592
 core saturation in, 575-576
 efficiency of, 64
 electrostatic shielding in, 578
 heating effect in, 575

- Transformers, losses in, 61-63
 power ratios of, 64-65
 regulation of, 62-63, 577-578
 three-phase, 590-592
- Transmission line antenna (TL), 562-567
- Transmission lines, 282-291
- Transmitter ratings, broadcast, 392-395
- Transmitter troubles, 448-452, 458-460, 467-469
- Transmitters, aircraft, 456-460, 461-471
 broadcast, 396-420
 care and maintenance of, 447-448
 marine, 421-453
- Transmitting-circuit principles, 155-206
- Transmitting tubes, 721-726, 728-729
- Trap circuits, 193, 198-200, 249-250
- Troubles, direction finder, 560
 generator, 651-653
 motor, 649-653
 receiver, 498-499, 560
 rectifier, 615-616
 storage battery, 696-697
 transmitter, 448-452, 458-460, 467-469
- Tube (*see* Vacuum tube)
- Tuned circuits, theory of, 129-131, 169-175, 179-186, 241-250, 287-291
- Tuned-plate, tuned-grid oscillator, 154
- Tungar charger, 710-712
- Tuning, broadcast transmitter, 399-403
- U
- Ultra-high-frequency, antennas, 293-294
 oscillators, 204-205, 474
 receivers, 253, 518
 systems, 201-206
 transmitters, 471-474
 waves, 275-277
- U. S. Coast Guard (*see* Coast Guard)
- U. S. Lighthouse service, charts, 539-541
- Unity power factor, adjustment, 403
 (*See also* Resonance)
- V
- Vacuum contact relay, 655-666
- Vacuum tube, alternating-current-operated, 124-127
 amplifier, 214-232
 (*See also* Power amplifiers)
 characteristics of, 121-124
 charts, 188, 189, 192, 209, 714-731
 detector, 209-214, 233-235
 diode, 233-235
 efficiency of, 132
- Vacuum tube, oscillators, 147-154
 peak voltmeter, 411
 pentode type of, 225-227
 screen-grid type of, 207-209
 theory of the, 116-121
 triode type of, 116-122
 voltmeter, 348, 377-380, 404-406
 water-cooled, 419-420
- Variable- μ tube, 209
- Vectors, 84-85, 88-96, 101-109, 112-113
- Velocity microphone, 330-331
- Voltage, dividers, 608-613
 feed antenna, 287
 inverse-peak, 583-584
 r.m.s., 87-88
 regulation, 375, 577-578
 storage cell, 683, 701
 (*See also* Electromotive force)
- Voltmeters, 659-661, 663-665
 vacuum tube, 348, 377-380, 404-406
- Volume controls, automatic, 235-241
 broadcast, 342, 347-357, 368
- Volume indicator, 348, 377-380, 404, 406
- W
- Water-cooling system, vacuum tube, 419-420
- Watt, 26-29
- Wattmeter, 668
 (*See also* Ampere-hour meter)
- Wave bands, 155
- Wave length, 265-266
- Wave-length-frequency conversion table, 732-735
- Wave meter, 304
- Wave polarization, 276-277
- Wave propagation, 265-282
- Wave reflections, 282-283
 (*See also* Sky waves)
- Wave traps, 249
- Waves, night effect on radio, 527, 560-562
 standing, 264, 282-283, 293
- Weather signals, 743
- Western Electric, aircraft receivers, 499-501
 aircraft transmitters, 456-469
 broadcast transmitters, 396-410
 frequency monitor, 407-410
 tube charts, 727-729
- Work, 26
- Wye (Y) transformer connections, 111, 589

THREE-PLACE LOGARITHMIC TABLES

THREE-PLACE LOGARITHMIC TABLES
Complete Logarithms of Numbers 1 to 100

N.	log	N.	log	N.	log	N.	log	N.	log
1	0.000	21	1.322	41	1.613	61	1.785	81	1.908
2	0.301	22	1.342	42	1.623	62	1.792	82	1.914
3	0.477	23	1.362	43	1.633	63	1.799	83	1.919
4	0.602	24	1.380	44	1.643	64	1.806	84	1.924
5	0.699	25	1.397	45	1.653	65	1.813	85	1.929
6	0.778	26	1.415	46	1.663	66	1.820	86	1.934
7	0.845	27	1.431	47	1.672	67	1.826	87	1.940
8	0.903	28	1.447	48	1.681	68	1.833	88	1.944
9	0.954	29	1.462	49	1.690	69	1.839	89	1.949
10	1.000	30	1.477	50	1.699	70	1.845	90	1.954
11	1.041	31	1.491	51	1.708	71	1.851	91	1.959
12	1.079	32	1.505	52	1.716	72	1.857	92	1.964
13	1.114	33	1.518	53	1.724	73	1.863	93	1.968
14	1.146	34	1.531	54	1.732	74	1.869	94	1.973
15	1.176	35	1.544	55	1.740	75	1.875	95	1.978
16	1.204	36	1.556	56	1.748	76	1.881	96	1.982
17	1.230	37	1.568	57	1.756	77	1.886	97	1.987
18	1.255	38	1.580	58	1.763	78	1.892	98	1.991
19	1.279	39	1.591	59	1.771	79	1.898	99	1.996
20	1.301	40	1.602	60	1.778	80	1.903	100	2.000

Mantissas of Numbers 1 to 2,009

N.	0	1	2	3	4	5	6	7	8	9
10	000	004	009	013	017	021	025	029	033	037
11	041	045	050	053	057	061	064	068	072	076
12	079	083	086	090	093	097	100	104	107	111
13	114	117	121	124	127	130	134	137	140	143
14	146	149	152	155	158	161	164	167	170	173
15	176	179	182	185	188	190	193	196	199	201
16	204	207	210	212	215	217	220	223	225	228
17	230	233	236	238	241	243	246	248	250	253
18	255	258	260	262	265	267	270	272	274	276
19	279	281	283	286	288	290	292	294	297	299
20	301	303	305	307	310	312	314	316	318	320
21	322	324	326	328	330	332	334	336	338	340
22	342	344	346	348	350	352	354	356	358	360
23	362	364	365	367	369	371	373	375	377	378
24	380	382	384	386	387	389	391	393	394	396
25	398	400	401	403	405	407	408	410	412	413
26	415	417	418	420	422	423	425	427	428	430
27	431	433	435	436	438	439	441	442	444	446
28	447	449	450	452	453	455	456	458	459	461
29	462	464	465	467	468	470	471	473	474	476
30	477	479	480	481	483	484	486	487	489	490
31	491	493	494	496	497	498	500	501	502	504
32	505	507	508	509	511	512	513	515	516	517
33	519	520	521	522	524	525	526	528	529	530
34	531	533	534	535	537	538	539	540	542	543
35	544	545	547	548	549	550	551	553	554	555
36	556	558	559	560	561	562	563	565	566	567
37	568	569	571	572	573	574	575	576	577	579
38	580	581	582	583	584	585	587	588	589	590
39	591	592	593	594	595	597	598	599	600	601
40	602	603	604	605	606	607	609	610	611	612
41	613	614	615	616	617	618	619	620	621	622
42	623	624	625	626	627	628	629	630	631	632
43	633	634	635	636	637	638	639	640	641	642
44	643	644	645	646	647	648	649	650	651	652
45	653	654	655	656	657	658	659	660	661	662
46	663	664	665	666	667	667	668	669	670	671
47	672	673	674	675	676	677	678	679	679	680
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49	690	691	692	693	694	695	695	696	697	698

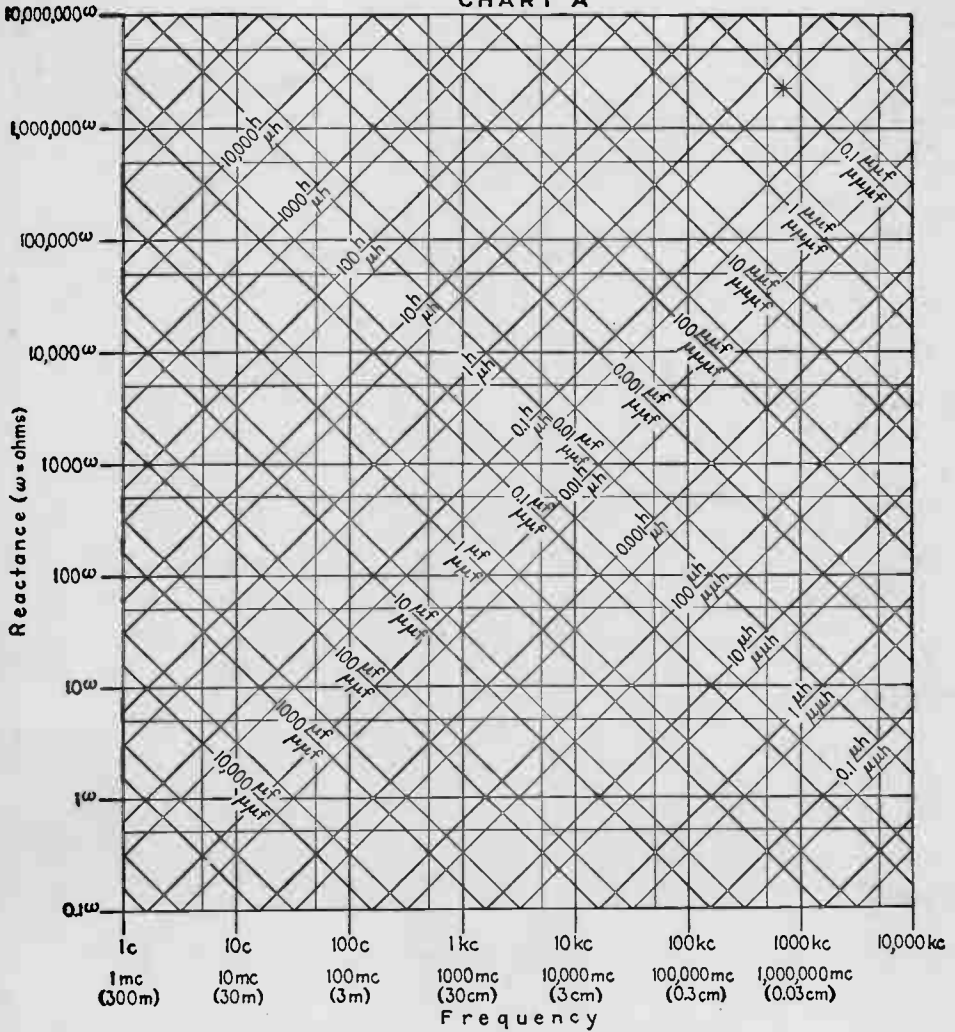
Mantissas of Numbers 1 to 2,009

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51	708	709	710	711	712	713	714	715	716	717
52	718	719	720	721	722	723	724	725	726	727
53	728	729	730	731	732	733	734	735	736	737
54	738	739	740	741	742	743	744	745	746	747
55	748	749	750	751	752	753	754	755	756	757
56	758	759	760	761	762	763	764	765	766	767
57	768	769	770	771	772	773	774	775	776	777
58	778	779	780	781	782	783	784	785	786	787
59	788	789	790	791	792	793	794	795	796	797
60	798	799	800	801	802	803	804	805	806	807
61	808	809	810	811	812	813	814	815	816	817
62	818	819	820	821	822	823	824	825	826	827
63	828	829	830	831	832	833	834	835	836	837
64	838	839	840	841	842	843	844	845	846	847
65	848	849	850	851	852	853	854	855	856	857
66	858	859	860	861	862	863	864	865	866	867
67	868	869	870	871	872	873	874	875	876	877
68	878	879	880	881	882	883	884	885	886	887
69	888	889	890	891	892	893	894	895	896	897
70	898	899	900	901	902	903	904	905	906	907
71	908	909	910	911	912	913	914	915	916	917
72	918	919	920	921	922	923	924	925	926	927
73	928	929	930	931	932	933	934	935	936	937
74	938	939	940	941	942	943	944	945	946	947
75	948	949	950	951	952	953	954	955	956	957
76	958	959	960	961	962	963	964	965	966	967
77	968	969	970	971	972	973	974	975	976	977
78	978	979	980	981	982	983	984	985	986	987
79	988	989	990	991	992	993	994	995	996	997
80	998	999	000	001	002	003	004	005	006	007
81	008	009	010	011	012	013	014	015	016	017
82	018	019	020	021	022	023	024	025	026	027
83	028	029	030	031	032	033	034	035	036	037
84	038	039	040	041	042	043	044	045	046	047
85	048	049	050	051	052	053	054	055	056	057
86	058	059	060	061	062	063	064	065	066	067
87	068	069	070	071	072	073	074	075	076	077
88	078	079	080	081	082	083	084	085	086	087
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90	098	099	100	101	102	103	104	105	106	107
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92	118	119	120	121	122	123	124	125	126	127
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117	368	369	370	371	372	373	374	375	376	377
118	378	379	380	381	382	383	384	385	386	387
119	388	389	390	391	392	393	394	395	396	397
120	398	399	400	401	402	403	404	405	406	407
121	408	409	410	411	412	413	414	415	416	417
122	418	419	420	421	422	423	424	425	426	427
123	428	429	430	431	432	433	434	435	436	437
124	438	439	440	441	442	443	444	445	446	447

Mantissas of Numbers 1 to 2,009

N.	0	1	2	3	4	5	6	7	8	9
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126	100	101	101	101	102	102	102	103	103	103
127	104	104	104	105	105	106	106	106	107	107
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137	137	137	137	138	138	138	139	139	139	140
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142	152	153	153	153	154	154	154	154	155	155
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145	161	162	162	162	163	163	163	163	164	164
146	164	165	165	165	166	166	166	166	167	167
147	167	168	168	168	168	169	169	169	170	170
148	170	171	171	171	171	172	172	172	173	173
149	173	173	174	174	174	175	175	175	176	176
150	176	176	177	177	177	178	178	178	178	179
151	179	179	180	180	180	180	181	181	181	182
152	182	182	182	183	183	183	184	184	184	184
153	185	185	185	186	186	186	186	187	187	187
154	188	188	188	188	189	189	189	189	190	190
155	190	191	191	191	191	192	192	192	193	193
156	193	193	194	194	194	195	195	195	195	196
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160	204	204	205	205	205	205	206	206	206	207
161	207	207	207	208	208	208	208	209	209	209
162	210	210	210	210	210	211	211	211	212	212
163	212	212	213	213	213	214	214	214	214	215
164	215	215	215	216	216	216	216	217	217	217
165	217	218	218	218	219	219	219	219	220	220
166	220	220	221	221	221	221	222	222	222	222
167	223	223	223	223	224	224	224	225	225	225
168	225	226	226	226	226	227	227	227	227	228
169	228	228	228	229	229	229	229	230	230	230
170	230	231	231	231	231	232	232	232	232	233
171	233	233	234	234	234	234	235	235	235	235
172	236	236	236	236	237	237	237	237	238	238
173	238	238	239	239	239	239	240	240	240	240
174	241	241	241	241	242	242	242	242	243	243
175	243	243	244	244	244	244	245	245	245	245
176	246	246	246	246	246	247	247	247	247	248
177	248	248	248	249	249	249	249	250	250	250
178	250	251	251	251	251	252	252	252	252	253
179	253	253	253	254	254	254	254	255	255	255
180	255	256	256	256	256	256	257	257	257	257
181	258	258	258	258	259	259	259	259	260	260
182	260	260	261	261	261	261	262	262	262	262
183	262	263	263	263	263	264	264	264	264	265
184	265	265	265	266	266	266	266	266	267	267
185	267	267	268	268	268	268	269	269	269	269
186	270	270	270	270	271	271	271	271	271	272
187	272	272	272	273	273	273	273	273	274	274
188	274	274	275	275	275	275	276	276	276	276
189	276	277	277	277	277	278	278	278	278	279
190	279	279	279	279	280	280	280	280	281	281
191	281	281	281	282	282	282	282	283	283	283
192	283	284	284	284	284	284	285	285	285	285
193	286	286	286	286	286	287	287	287	287	287
194	288	288	288	288	289	289	289	289	290	290
195	290	290	290	291	291	291	291	292	292	292
196	292	292	293	293	293	293	294	294	294	294
197	294	295	295	295	295	296	296	296	296	296
198	297	297	297	297	298	298	298	298	298	299
199	299	299	299	300	300	300	300	300	301	301
200	301	301	301	302	302	302	302	303	303	303

CHART A



NOTE: Heavy lines = 1 coordinate; light lines = 5 coordinates

See Appendix A for directions for use of this chart.

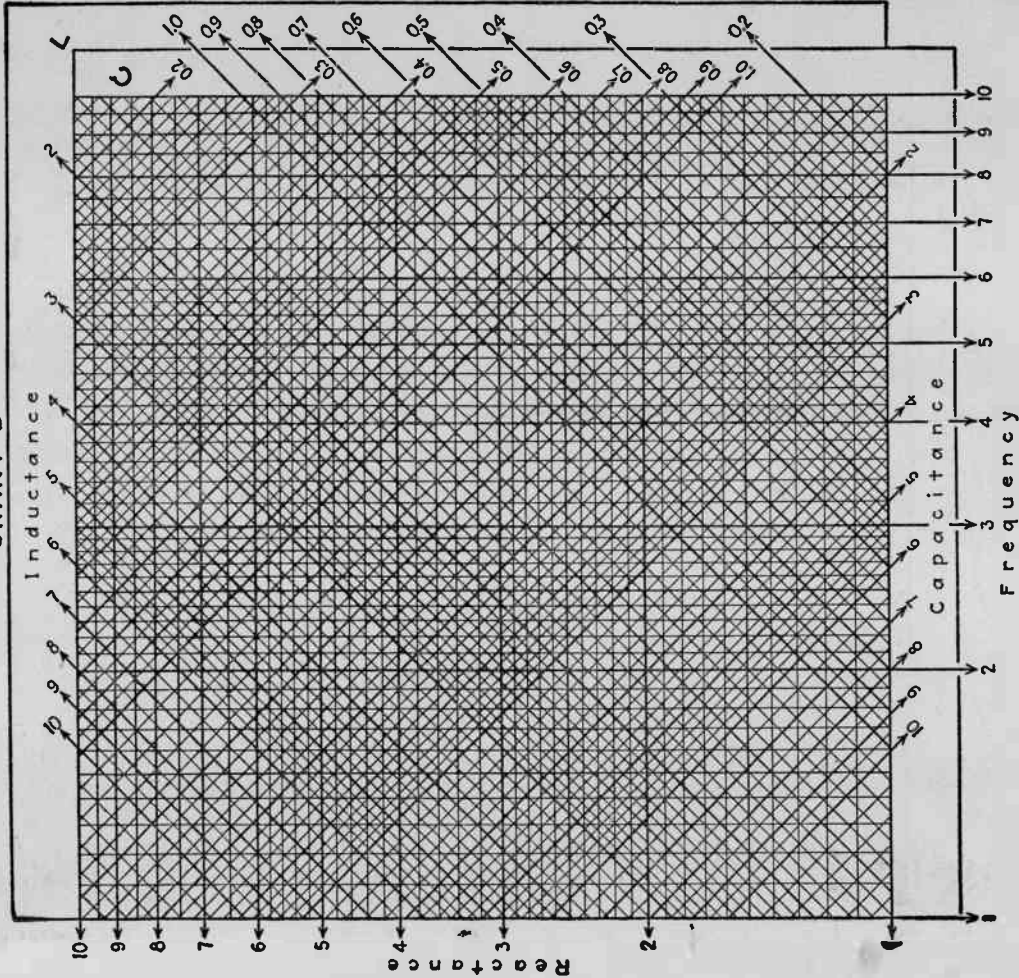
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100

CONVERSION TABLE

To convert	to	multiply by
Amperes	Microamperes	1,000,000 (10^6)
Amperes	Milliamperes	1,000 (10^3)
Centimeters	Henrys	0.000,000,001 (10^{-9})
Centimeters	Inches	0.3937
Centimeters	Meters	0.01 (10^{-2})
Centimeters	Microhenrys	0.001 (10^{-3})
Centimeters	Millihenrys	0.000,001 (10^{-6})
Centimeters	Micro-microfarads	1.1124
Cycles	Kilocycles	0.001 (10^{-3})
Cycles	Megacycles	0.000,001 (10^{-6})
Farads	Microfarads	1,000,000 (10^6)
Farads	Micro-microfarads	1,000,000,000,000 (10^{12})
Feet	Meters	0.305
Henrys	Centimeters	1,000,000,000 (10^9)
Henrys	Microhenrys	1,000,000 (10^6)
Henrys	Millihenrys	1,000 (10^3)
Horsepower	Watts	746
Inches	Centimeters	2.54
Joules	Ergs	10,000,000 (10^7)
Kilocycles	Cycles	1,000 (10^3)
Kilovolts	Volts	1,000 (10^3)
Kilowatts	Horsepower	1.34
Kilowatts	Joules per second	1,000 (10^3)
Kilowatts	Watts	1,000 (10^3)
Meters	Feet	3.28
Mhos	Micromhos	1,000,000 (10^6)
Microamperes	Amperes	0.000,001 (10^{-6})
Microfarads	Farads	0.000,001 (10^{-6})
Microhenrys	Henrys	0.000,001 (10^{-6})
Microhenrys	Centimeters	1,000 (10^3)
Micromhos	Mhos	0.000,001 (10^{-6})
Microvolts	Volts	0.000,001 (10^{-6})
Microwatts	Watts	0.000,001 (10^{-6})
Micro-microfarads	Farads	0.000,000,000,001 (10^{-12})
Micro-microfarads	Microfarads	0.000,001 (10^{-6})
Milliamperes	Amperes	0.001 (10^{-3})
Millihenrys	Centimeters	1,000,000 (10^6)
Millihenrys	Henrys	0.001 (10^{-3})
Millivolts	Volts	0.001 (10^{-3})
Milliwatts	Watts	0.001 (10^{-3})
Ohms	Megohms	0.000,001 (10^{-6})
Ohms	Microhms	1,000,000 (10^6)
Volts	Microvolts	1,000,000 (10^6)
Volts	Millivolts	1,000 (10^3)
Watts	Kilowatts	0.001 (10^{-3})
Watts	Microwatts	1,000,000 (10^6)
Watts	Milliwatts	1,000 (10^3)

CHART B



See Appendix A for directions for use of this chart.

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*James B. Holliday
Ballups Island, Moos.*

INDEX TO CHAPTERS FOR QUICK REFERENCE

Chapter	Page
I. Direct-current Electricity and Magnetism.....	I
II. Alternating-current Electricity.....	58
III. Introduction to Vacuum Tubes.....	116
IV. Transmitting-circuit Principles.....	155
V. Receiving-circuit Principles.....	207
VI. Antennas and Wave Propagation.....	261
VII. Studio Acoustics and Apparatus.....	309
VIII. Control-room Equipment and Operation.....	345
IX. Broadcast Transmitters.....	391
X. Communication Transmitters.....	421
XI. Radio Receivers.....	475
XII. Radio Aids to Navigation.....	523
XIII. Rectifier Units.....	575
XIV. Dynamo-electric Machinery and Meters.....	617
XV. Storage Batteries.....	671
Appendix A (Technical Information).....	713
Appendix B (Operating Information).....	739

