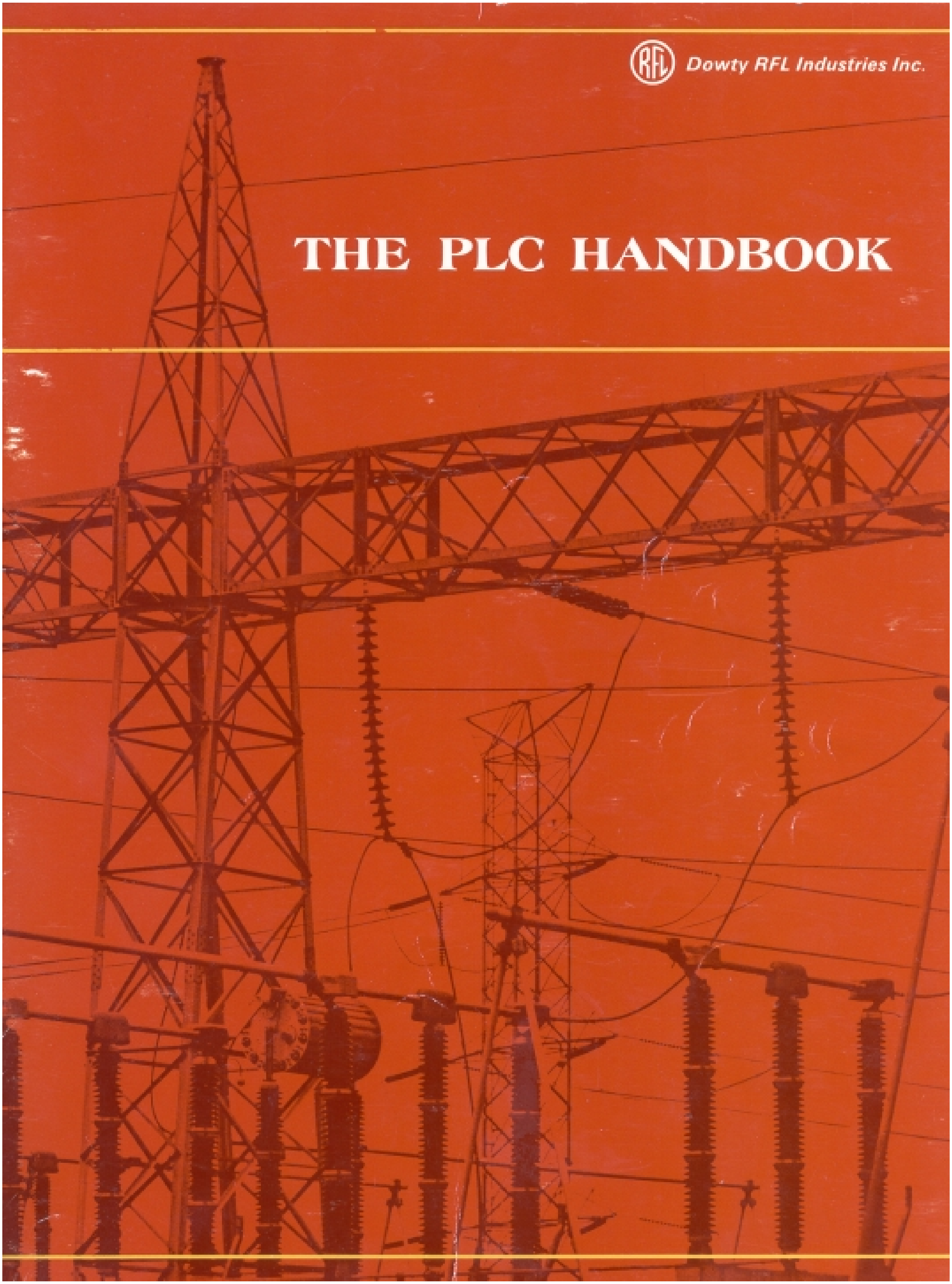
Dowty RFL Industries Inc.

THE PLC HANDBOOK



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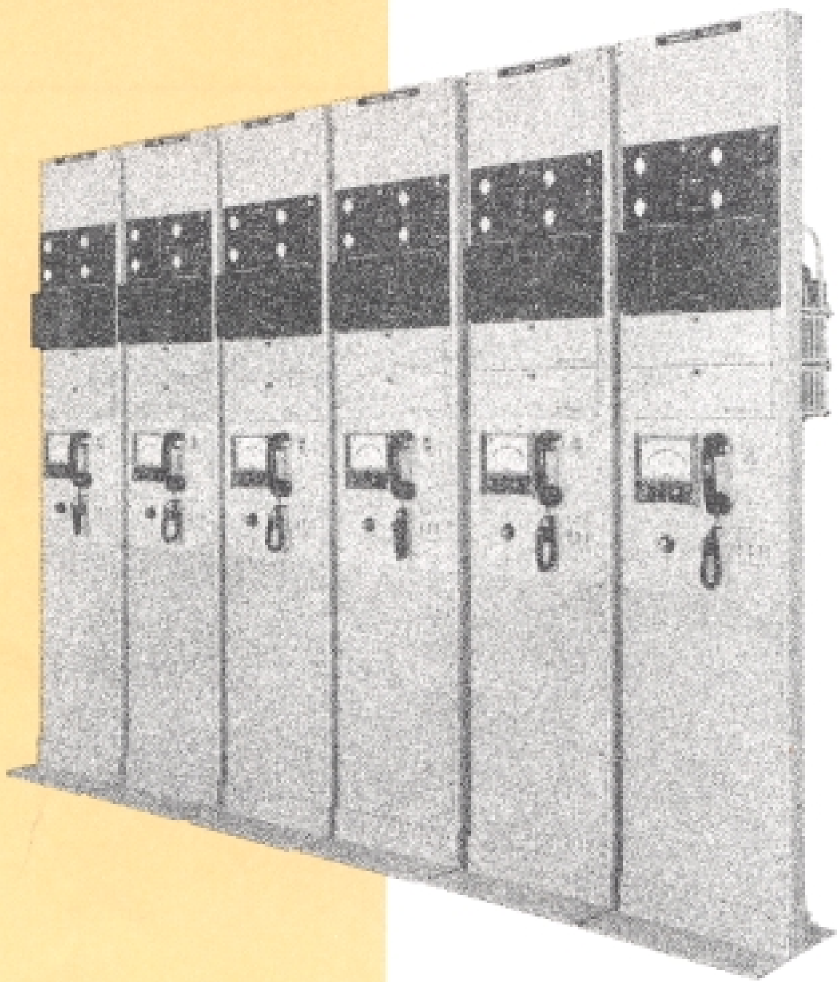

Frontispiece: Six of eighteen stations supplied by RFL for a large PLC system.

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FUNDAMENTALS OF PLC

INTRODUCTION

Powerline carrier (PLC) is the brief term used to denote the entire process of communication which uses high-voltage powerlines as the means for transmission. Powerlines provide a reliable link because of their unusually rugged construction. They also offer the advantage of freedom from restrictions imposed by common-carrier regulations, as well as lower cost. As compared to microwave and radio links, they offer low cost and relative freedom from fading and weather conditions.

PLC communications will typically be in the form of speech and/or tones (1) for telegraph, telemetering, telecontrol, higher speed data applications, and line protection. The latter function will often be the most important and critical application for the powerline-carrier system.

The large magnitude of difference between the voltages and currents of power transmission and the much lower powers practical for communication signals is offset by the large difference in frequency between that of the power transmitted and of the carrier band used for communication, typically between 30 and 500 kHz. The frequency separation is accomplished by coupling capacitors and line traps or inductors, in conjunction with coupling devices or line-tuning units.

A PLC channel includes the signal path from the transmitting equipment at one terminal, through the tuning equipment at the receiving end, and into the receiving terminal. In duplex (bidirectional) operation, a signal is sent over the same path in the opposite direction and on a different frequency.

In planning a channel, consideration must be given to the need for confining the signals to a desired path and to excluding unwanted signals from it. Such functions are achieved through use of line traps, coupling capacitors, and tuners.

(1) Voice-frequency-carrier signals, modulated either AM (usually on-off) or by frequency-shift keying (FSK) are universally used for combining many different signals on a single audio band by frequency multiplexing. The musical tone heard when these signals are audibly monitored causes them to be referred to as "tones", either singly or in multiple.

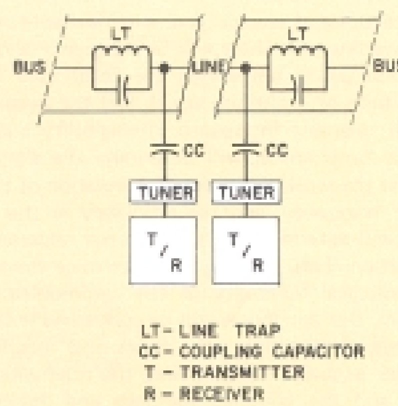

Figure 1.1. Typical PLC channel.

A typical PLC channel is outlined in Figure 1.1. A transmitter signal from one end passes through a tuner which, in conjunction with the coupling capacitor, provides a low-impedance path between the transmitter and the line.

OTHER APPLICATIONS

These notes deal specifically with the use of the RFL Series 65 Single-Sideband Carrier System as a communication terminal in systems in which a high-voltage powerline is the medium of transmission.

It may be noted, however, that all terminal equipment of the Series 65 is equally applicable to cases where other transmission media are used. These include insulated shield wires, openwire line (as on railroad rights of way), coaxial cable, microwave and radio systems. In such cases, the most significant differences are usually found in the means for coupling the terminal equipment to the transmission medium and in the choice of frequency plan. Though not discussed here, RFL has experience in and equipment for such applications.

The tuner contains an impedance-matching transformer, so that the terminal equipment and the line each see the proper impedance. The line trap provides a high impedance at the operating frequencies to prevent loss of carrier-signal power into the low-impedance station bus. Similarly, the line trap at the far end confines the signal power to its intended path, and it blocks noise and other interference which may be on the power bus.

Included in the definition of PLC are other transmission arrangements in which power cables, cables in series with overhead powerlines, or insulated ground (shield) wires are used as transmission paths. Various coupling methods are used for these combinations. Insulated shield wires, however, are not normally recommended for protective relaying.

Electrical noise of several types will inevitably be present on the powerline, and there will be frequency components of this noise which fall within the band used for PLC. The magnitude of this noise signal, and the signal-to-noise ratio (S/N) tolerable for speech intelligibility and reliable tone-circuit functioning, will determine the signal power necessary at the receiving station. Attenuation of the line at the carrier frequency, and the efficiency of the coupling employed will determine the signal power necessary at the sending station. This, in turn, will determine the economic and technological feasibility of the communication link. Accordingly, this handbook will be concerned with, among other things, discussion of the types and magnitudes of noise to be expected on the line, the attenuation to be expected, and the coupling methods and devices to be employed. There will also be a discussion of equipment used to generate and receive signals, voice-frequency carrier equipment for data and control functions, and line-protection equipment, all available from RFL.

MODE OF TRANSMISSION

Historically, communication signals carried over high-voltage powerlines have used nearly every form of modulation known. From this experience, modern practice has resolved upon the almost exclusive use of amplitude-modulated, single-sideband suppressed-carrier systems, generally abbreviated SSB. The system combines a number of desirable features, among which are economy of bandwidth and of transmitter power. The RFL Series 6515 System is an SSB system although, of course, the multiplexed channels which it carries may contain information encoded in other forms.

The conservation of bandwidth compared to double-sideband AM or FM not only increases the possible signal density of the system, but it also reduces the noise present within each channel. The concentration of available power in intelligence reduces the overall power requirement of the system.

FREQUENCY RANGES

The Model 65 MOD is the SSB modulator of the Series 65 PLC System. By choice of oscillator crystals, the carrier frequency of the output signal may be any one of the fourteen frequencies between 4 and 60 kHz which are set equally 4 kHz apart. In addition, the output signal may be selected to be either upper or lower sideband by choosing an output filter of appropriate type and frequency range. The Model 65 DEMOD is the complementary unit at the receiving terminal. These modules are the basic elements of a system capable of operating anywhere in the carrier-frequency range between 4 and 60 kHz.

The low-frequency range below 20 kHz is often used for openwire-line systems and others; but for PLC systems the limitations in the coupling elements make the low-frequency range impractical.

The maximum of fourteen channels below 60 kHz also can be limiting because of such things as conflicting requirements for frequency assignments, low-frequency noise, need for more channels, or for the superior performance often available at higher frequencies. The need for higher carrier frequencies is met, in the Series 65 System, with the Model 65 UPCONV Upconverter. This module accepts the output of the Model 65 MOD, usually in the range between 20 and 30 kHz, and translates it to a high-frequency position for transmission. The carrier frequency may be anywhere in the range between 60 and 512 kHz. The Model 65 DNCONV Downconverter is the complementary receiving module. Its output, anywhere in the range between 4 and 60 kHz, is fed to a Model 65 DEMOD, for conversion to the audio band.

When selecting the operating frequencies for a PLC system, the performance specification of the equipment and the capabilities of the transmission medium used may not be the only limitation with respect to choice of operating frequency. The designer of the system should also be certain that the proposed frequency plan conforms to whatever administrative limitations may exist, with respect to frequency bands or sub-bands, in the country or geographic area in which operation is planned. In this connection, it should be remembered that lower and upper sidebands extend, respectively, beyond the lower and upper carrier frequencies of any bands selected.

FREQUENCY TRANSLATION

Where a single voice channel is to be translated to the high-frequency portion of a PLC frequency band, using single-sideband techniques, the translation is commonly accomplished in two steps as illustrated on Figures 1.2 through 1.9. Figure 1.2 shows the frequency spectrum of a voice-frequency (v-f) channel containing speech in the band from 300 to 3400 Hz, and a signaling tone at approximately 3825 Hz. The speech and tone spectra have been limited by highpass and lowpass filters for speech, and by a bandpass filter for the tone. Figure 1.3 shows the spectrum after modulation with a 24 kHz carrier. The carrier frequency has been suppressed by the balanced modulator, and is shown here 40 dB below the voice level. Both upper and lower sidebands are present at this point. Figure 1.4 shows the amplitude characteristic of a lower-sideband bandpass filter used as an IF filter, and Figure 1.5 shows the filtered spectrum with the carrier further suppressed and the unwanted sideband removed.

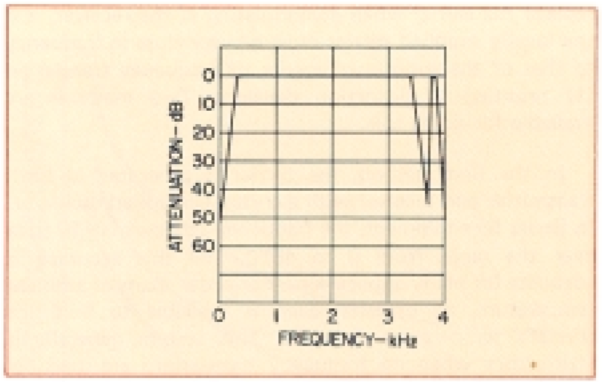

Figure 1.2. Spectrum of typical v-f channel.

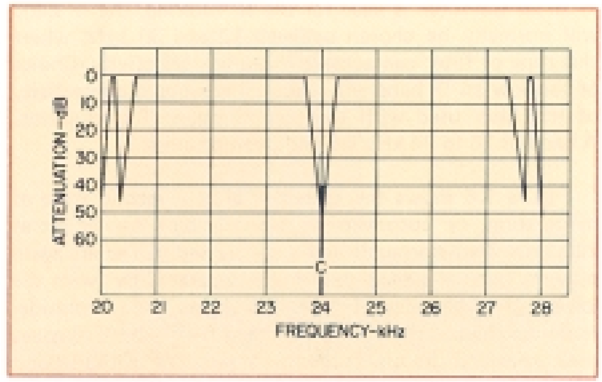

Figure 1.3. Typical v-f channel after modulation with a 24-kHz carrier.

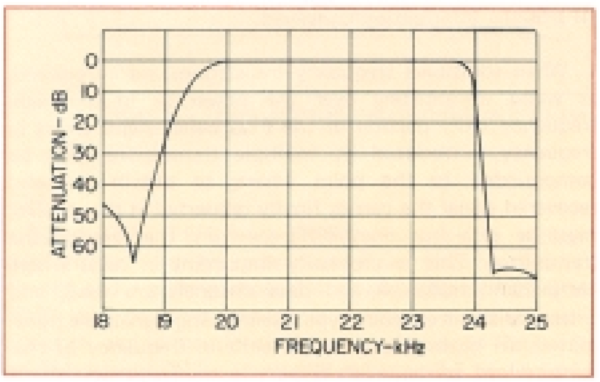

Figure 1.4. Attenuation characteristic of lower-sideband filter used as IF filter for signal in Figure 1.3.

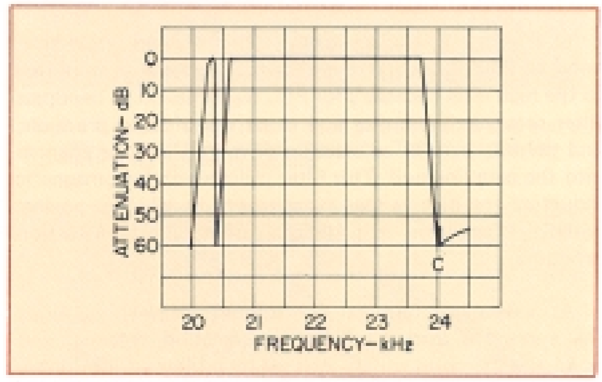

Figure 1.5. Spectrum of signal in Figure 1.3 following transmission through bandpass filter with characteristic shown in Figure 1.4.

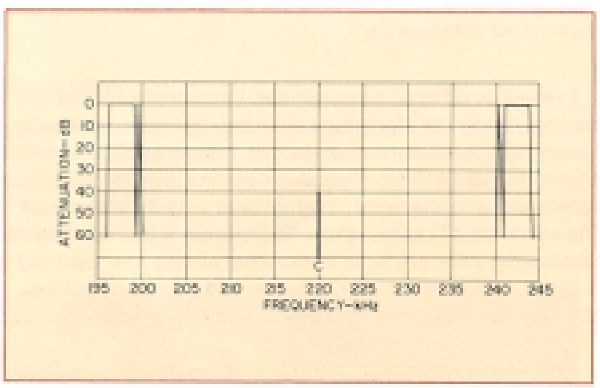

Figure 1.6. Signal spectrum after upconversion.

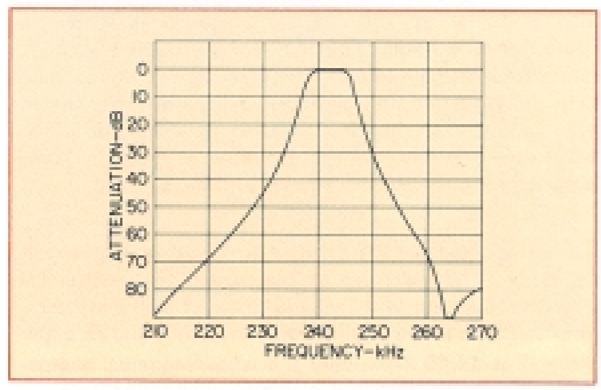

Figure 1.7. Attenuation characteristic of bandpass filter used to pass the upper sideband, 240-244 kHz.

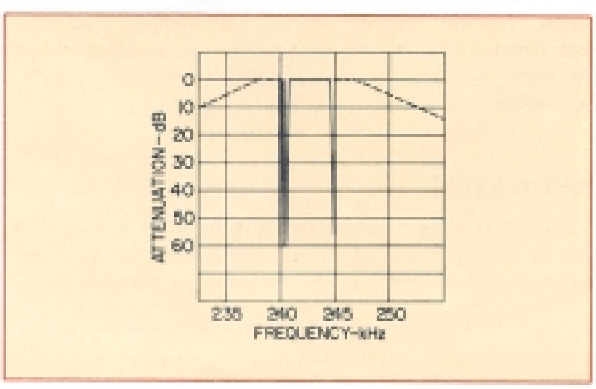

Figure 1.8. Spectrum of retained sideband at 240-244 kHz.

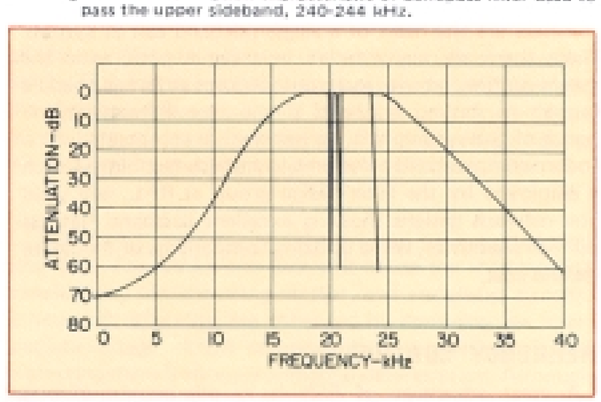

Figure 1.9. Spectrum of signal in Figure 1.8 after downconversion, using a carrier at 220 kHz.

If an LC filter is used for the IF filtering, the IF band will normally be chosen between 12 and 30 kHz, where this type of filter can achieve maximum selectivity. Choice of too low an IF band will require increasing the selectivity of the filter used with the second stage of modulation. A band of 20 to 24 kHz is a good compromise.

Figure 1.6 shows the spectrum after a second stage of modulation, or upconversion, when using a carrier set at 220 kHz. Two sidebands and a suppressed carrier are again present, but the sidebands are now separated by twice the lowest IF frequency. Figure 1.7 shows the amplitude-response characteristic of a bandpass filter which removes the carrier and the unwanted lower sideband of the second stage of modulation. Because of the separation of the side-bands, this filter need not be as selective as the IF filter.

In Figure 1.8, the spectrum of the retained upper side-band at 240 to 244 kHz is shown. This band is amplified to the high level necessary for PLC transmission. A bandpass filter removes harmonics and other out-of-band products, and permits parallel connection of more than one channel into the coupling unit. This filter utilizes non-ferromagnetic inductors and high-voltage capacitors to handle high powers without breakdown or generation of nonlinear distortion products.

At a receiving terminal of the same channel, assuming the system is loaded with communication channels and noise, the spectrum will be selected by a filter with response identical to that of the upconverter's bandpass filter shown in Figure 1.7. When this band is down-converted, using a carrier at 220 kHz, a spectrum as shown in Figure 1.9 will result. This includes the desired channel, translated back again to the IF band from 20-24 kHz. There is also an upper sideband, generated by the down-conversion demodulator, extending from 460 to 464 kHz, but a low-pass filter, incorporated in the down-converter's filter, completely suppresses both this and unwanted noise.

Another lower-sideband filter, as shown in Figure 1.4, will restrict the spectrum to the desired IF band, and a final demodulation with a 24-kHz carrier will generate a lower sideband at the original voice frequency and an upper sideband at 44-48 kHz. Only the voice-frequency components will pass the final speech and signaling-tone filters.

Filters are the heart of a single-sideband carrier system. Where there are many filters in a signal path, as in the system outlined above, each must be kept as flat as possible through its desired passband if the overall frequency response of the system is to be held within acceptable limits. Modern computerized polynomial-synthesis techniques, such as employed by the filter-design group at RFL, will yield filter-network designs showing excellent passband flatness and high selectivity, while utilizing components of moderate size and cost.

FREQUENCY CONTROL

A great advantage of SSB is its elimination of one sideband and of the carrier. It is necessary, however, to replace the carrier when demodulating at the receiver, and the locally supplied carrier must be very close in frequency to that of the suppressed carrier or frequency translation (1) resulting in distortion develops. Two methods are available for doing this.

In the first method, the carrier is generated at both transmitter and receiver with a crystal-controlled oscillator. In Series 65 equipment, the frequency accuracy is ± 10 ppm over the range from 0 to 45°C, and this accuracy is adequate for many applications. For wider spans of ambient temperature, an optional oven is available to hold the crystal's temperature constant. This system generally is satisfactory when no frequency translations are used between sending and receiving terminals other than the SSB modulation and demodulation at intermediate frequency (IF) of the information conveyed.

When additional frequency translations are introduced, as when transmitting over the powerline in the high-frequency (RF) portion of the PLC band, slight errors in frequency, introduced by multiple translations, may be compounded to the point where, to obtain a useful recovered signal the carrier finally reinserted at the receiver must be at a frequency different from that used at the transmitter. This is especially important in cases where narrowband telegraph and data channels are used, and where variable frequency-type telemetering signals are transmitted. In both cases, a small shift in frequency of the reintroduced carrier, with respect to the frequency of the carrier suppressed at the sending terminal, can introduce a shift which is a large percentage of the bandwidth of the transmitted intelligence.

Frequency differences introduced by multiple translations can be preserved for demodulation by simultaneously transmitting a pilot tone of carrier frequency. This, then, suffers the same frequency displacement as the sideband signal, and so a precisely synchronized system is obtained. The economy of power of the SSB system is not sacrificed, because the pilot carrier is transmitted at very low level and is easily recovered at the receiver with a selective filter. It is subsequently raised to a level suitable for demodulation.

The pilot carrier may also be used as a reference level for automatic-gain-control (AGC) because it is always transmitted at a fixed level with respect to the level of the information channel. It is recommended that this second method for providing carrier at the receiver be used whenever multiple frequency translations are used in the system.

MULTIPLEXING

Multiplexing communication functions on a PLC link may be accomplished in the audio-frequency band, in the intermediate-frequency band, in the radio-frequency band,

(1) This parasitic translation should be distinguished from the deliberate translations which are a part of the modulation pattern.

or in all three. In addition to combining various speech and tone functions within a single four-kHz channel, it is common to combine several such channels into an intermediate-frequency (IF) group occupying the region between voice-frequency audio and the lowest PLC-frequency band. In the Series 65 system, this IF band extends from 4 to 60 kHz. Next, two or more separate groups, each in the IF band, may be multiplexed in the PLC-frequency band by using frequency translators with different carrier frequencies.

Although modern filter-design techniques will permit parallel connection of adjacent signal channels in the intermediate-frequency range for multiplexing, it is preferable to separate the channels into two groups with alternating frequency assignments. One group may contain the even-numbered channels while the other contains the odd-numbered channels, so that within each group there are unused guard bands which minimize adjacent-channel loading effects. The two groups are finally combined in a summing amplifier.

A power amplifier is used to boost the power of the translated group to the level necessary for transmission. The outputs of two or more such carrier-frequency amplifiers may be combined into a single line coupler, or line-tuning unit (LTU), by using high-power filters following the amplifiers.

"SPEECH–PLUS" SYSTEMS

The usual PLC channel has a bandwidth of 4 kHz. The standard audio channel extends from 300 to 3400 Hz, and this may be used entirely for voice. A good-quality voice channel may be obtained, however, with a bandwidth as narrow as 2 kHz. This suggests that the standard 4-kHz channel can carry other information in the frequency band above 2 kHz. This is the essence of the "speech-plus" system, in which filters are added to the channel to restrict the voice-frequency band and thereby render the balance of the channel's passband available for transmitting other intelligence. Hence the term speech-plus-data, or simply "speech plus". The frequency band above the cutoff frequency of the lowpass audio filter is often termed the superaudio band.

Figure 1.10 shows the passband and attenuation characteristics for typical lowpass filters used in voice circuits. It is clear that tradeoffs between quality of speech and quantity of data are possible. Figure 1.11 shows some typical applications when the cutoff frequency of the voice channel has been set at 2400 Hz. Above this frequency, 8 channels of 50-baud data, three channels of 100-baud data, or one channel of 200-baud data can be transmitted and received. These are, of course, in addition to the signaling at 3825 Hz.

Lowpass filters are normally included in modules of the Series 65 System, whether required for superaudio applications, or otherwise. When not used for speech-plus multiplexing, they are still useful for limiting the voice-frequency spectrum to eliminate noise and, thereby, improve the S/N ratio. Lowpass filters also prevent voice-frequency signals from interfering with signaling-tone circuits, and they eliminate spurious signals which could be present at frequencies above 4 kHz.

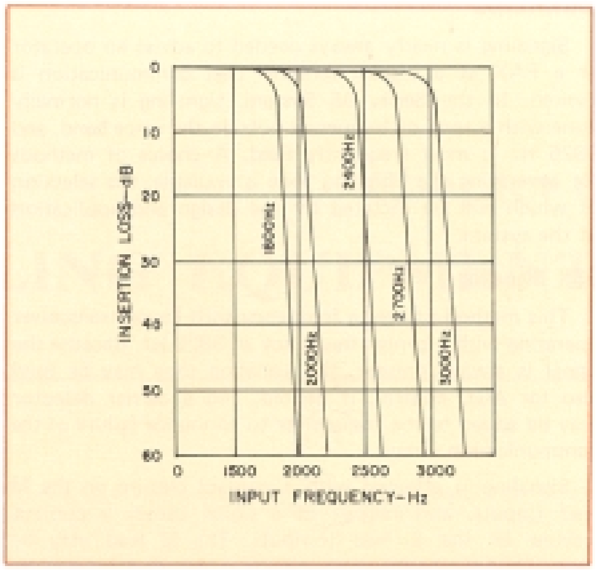

Figure 1.10. Typical lowpass-filter characteristics found in speech-plus systems.

Lowpass filters are included in the signal path as part of the modulators and demodulators. They permit addition of data and signaling after the voice path of the transmitting side, and they allow extraction of them prior to the voice path on the receiving side. They appear in the block diagrams of the Models 65 MOD and 65 DEMOD, given in Chapter 5, and their sequence in the signal path is shown in Figure 3.6.

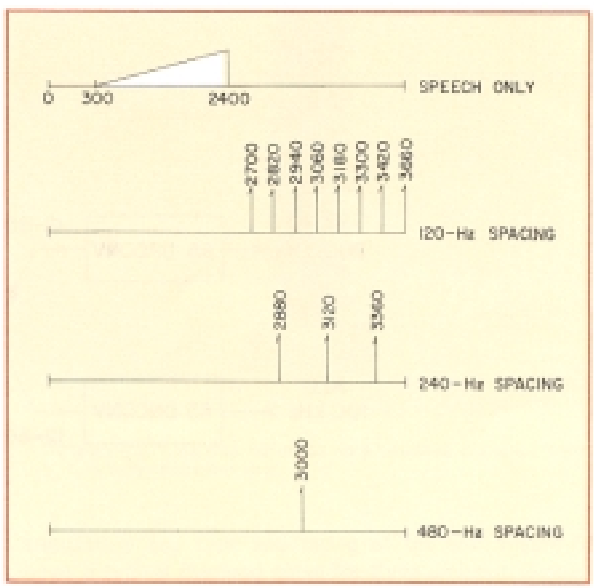

Figure 1.11. Typical data channels used with speech-plus systems. Not shown is the usual signaling channel at 3825 Hz.

SIGNALING

Signaling is nearly always needed to advise an operator or a PAX at a distant terminal that communication is desired. In the Series 65 System, signaling is normally done with a tone of frequency outside the voice band, and 3825 Hz is most frequently used. A choice of methods for generating the signaling tone is available, the selection of which will be dictated by the design and application of the system.

FSK Signaling

This method utilizes a frequency-shift-keyed transceiver operating with a center frequency of 3825 Hz. Because the signal is always present, the signaling tone may be used also for AGC control, if desired, and a carrier detector may be added to the transceiver to announce failure of the communication circuit.

Signaling is effected with a contact closure on the M lead (input), and receipt of a signal causes a contact closure on the E lead (output). The M lead may be controlled by a pushbutton, a pulse dialer, or other device. The signal from the relay output constituting the E lead can be used to operate a visual, audible, or other alerting device, or it may be extended to a local PBX at the receiving site where it may be used to originate pulse dialing over the local telephone system.

Orderwire System

When the PLC system is used to carry a Series 6850 Orderwire System, either or both of the two signaling methods available with the orderwire may be used.

When used with Model 6500 Selective Calling Telephone Unit both of the two signaling methods are used.

In one method, signaling is effected with an ON-OFF keying of a 3825-Hz signal. Operation of the E and M circuits is exactly the same as discussed earlier. Because ON-OFF keying is used, this signal cannot provide either AGC control or a carrier detector.

A second method of signaling, using the orderwire or SCTU systems, is in-band, voice-frequency dual-tone, multi-frequency (DTMF) signaling, by which up to 999 different stations may be selectively called, either jointly or severally, by pressing buttons on a keypad. Because the signals are in the voice band, the DTMF tones provide for easy extension of "dialing" through a PBX to a telephone network.

REPEATERS

Where repeaters are needed because of loss in a long line, because of loss in needed bypasses, or to provide a voice drop at an intermediate location, it is possible either to drop the carrier channels down in frequency to IF level, or to drop them all the way down to audio frequency. If no voice drop is required, it is less expensive to repeat at the IF level, since only one stage of frequency conversion is required and an entire group can be repeated with a single amplifier. Also, because of the lesser amount of equipment in a single path, the distortion introduced by the repeater will be less. In general, repeaters transmit the signal on a frequency different from the received frequency. A system repeating at IF level is outlined in Figure 1.12.

If voice drop is required, a complete IF demodulator and modulator for each voice-frequency channel of interest will be needed, but this will provide added selectivity and versatility. In such cases, the functional constituency of each channel also can be changed such as, for example, repeating the voice while removing certain tone functions and adding others at the repeater site. Either of two circuit arrangements is suitable for obtaining an audio drop.

In the system outlined on Figure 1.13, high-frequency signals are translated down to intermediate frequency and then translated as a group up to a different frequency in the RF band after passing through a four-way, four-wire bridge. Frequency assignments shown are only exemplary. Note that all signals pass through at intermediate frequency.

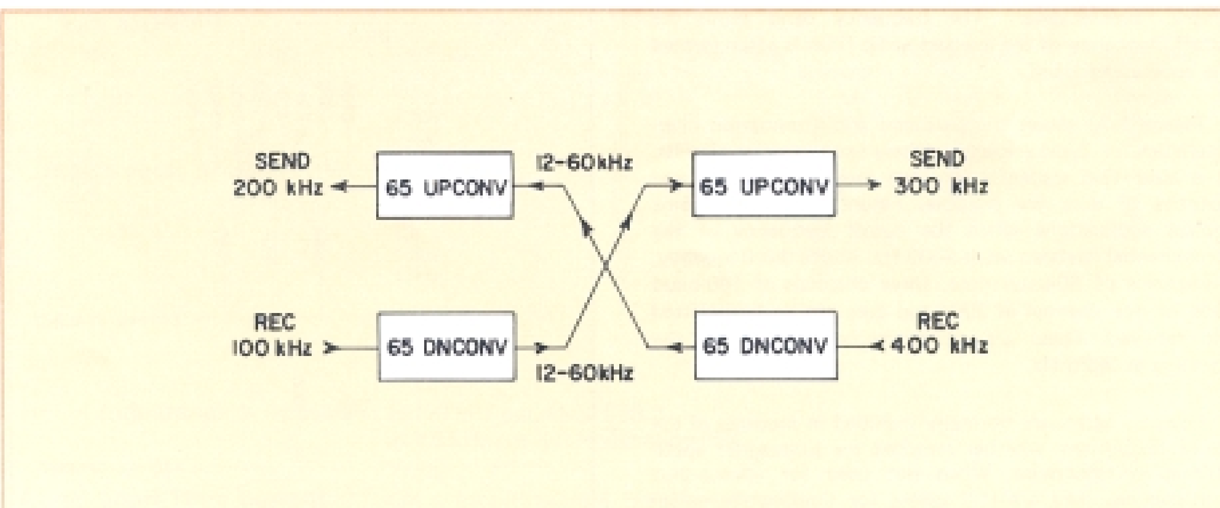

Figure 1.12. Duplex repeater without audio drop.

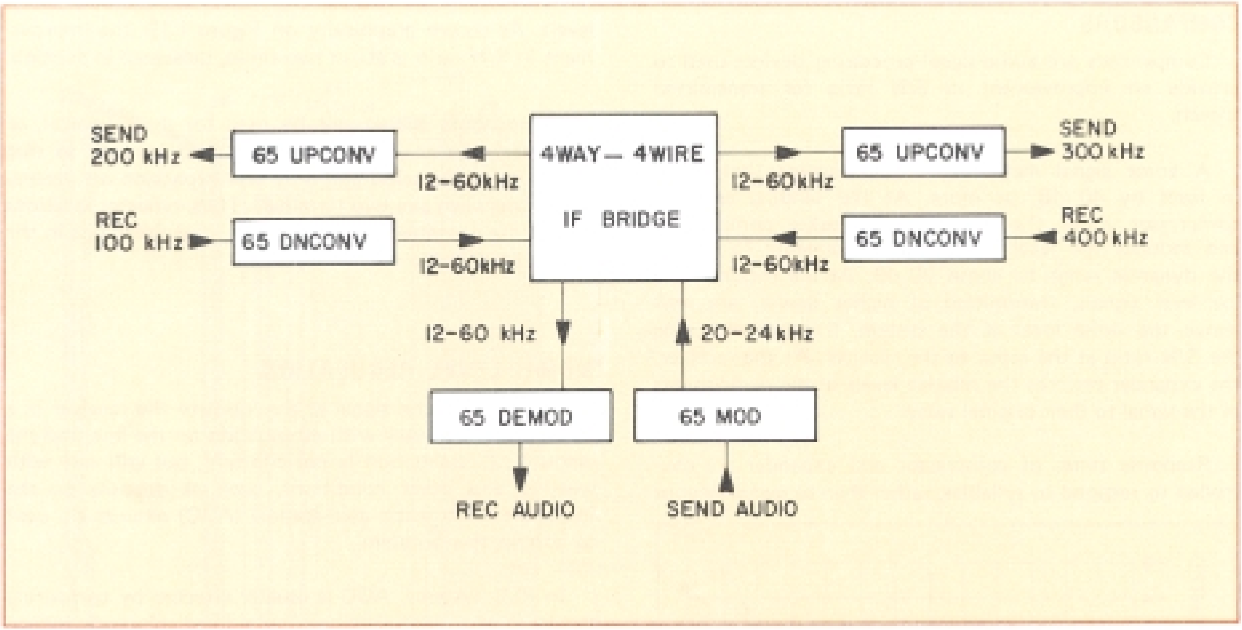

Figure 1.13. Duplex repeater with single audio drop, and with IF hybrid used to repeat also a group of other signals.

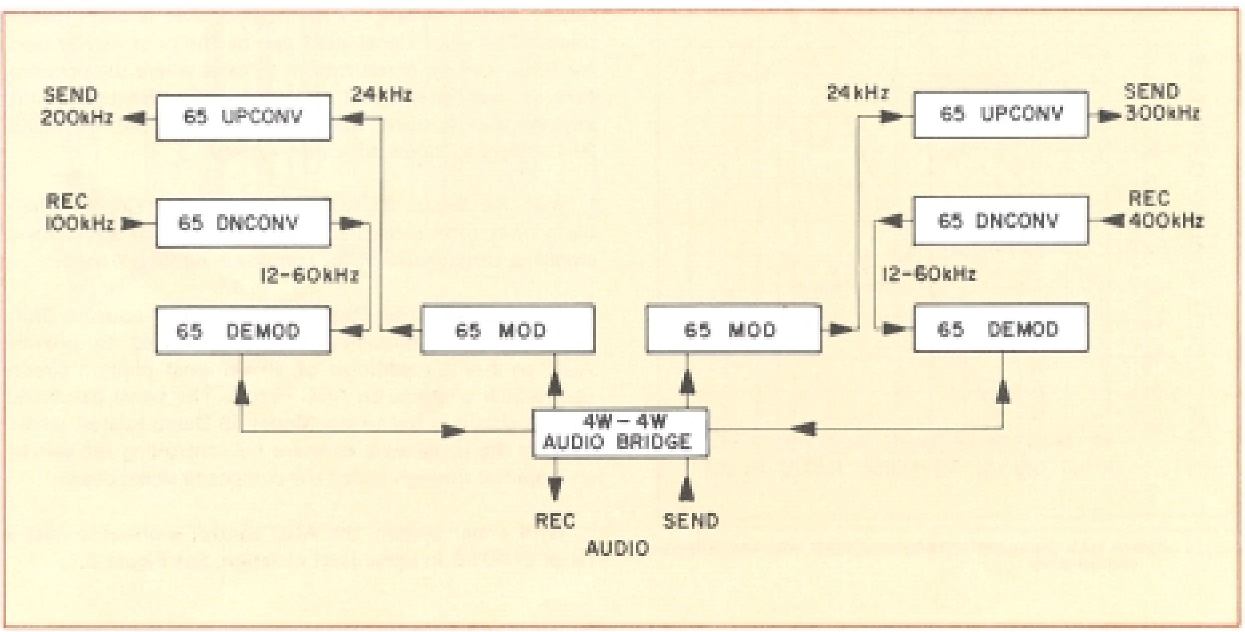

Figure 1.14. Duplex repeater with audio drop for every channel. A separate modulator or demodulator is required for every channel sent or received in either direction.

This arrangement is useful for a single audio drop because of the economy of equipment, although a loss at IF level of about 15 dB through the passive combiner may have to be made up elsewhere. Moreover, the loss through a passive combiner may be sidestepped, if needed, by using an active bridge, which may be simply assembled by using two Model 65 WB DUAL AMP circuit cards.

A second arrangement for obtaining an audio drop is shown in Figure 1.14. This system will find application where only one channel is carried and must be dropped from RF level, or where all channels must be dropped to the voice band. In either case a modulator or a demodulator will be required for each channel in each direction. Frequency assignments shown are only typical.

The performance requirements established for a PLC system will lead the designer to the most economical choice among the various repeater systems outlined.

COMPANDORS

Compandors are audio-signal-processing devices used to provide an improvement in S/N ratio for transmitted speech.

A voice signal may contain components that differ in level by 40 dB, or more. At the sending end the compressor raises the level of the weaker components and reduces the level of strong components to reduce the dynamic range to about 20 dB. At the receiver, the low-level signals, transmitted at higher power, are well above the noise level of the system, thereby improving the S/N ratio at the input to the receiver. At the receiver, the expander restores the relative levels of the components of the signal to their original value.

Response times of compressor and expander are controlled to respond to syllables, rather than to instantaneous levels. As shown graphically on Figure 1.15, the improvement in S/N ratio is about two times, measured in decibels.

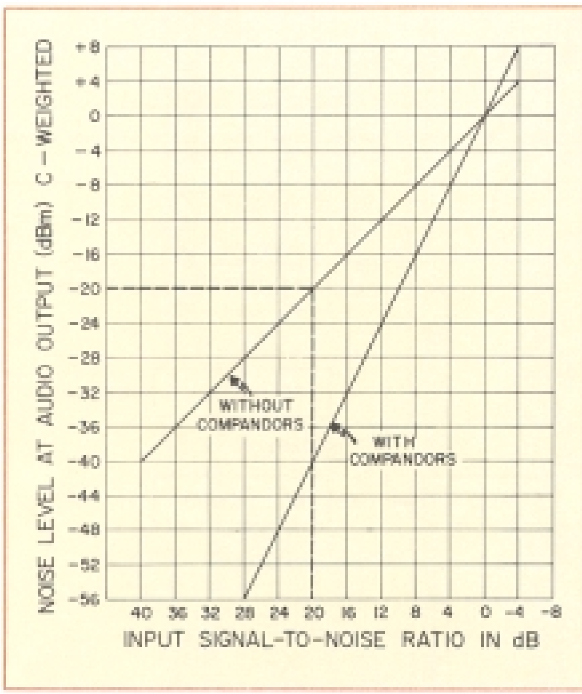

Figure 1.15. Noise performance of system with and without compandors.

Compandors should not be used for data, control, or telemetering channels. Compandors must be used so that only one compression and only one expansion are accomplished between any two terminals. Thus, repeater locations should use a compandor only in the local drop, not in the direct-repeat path.

SIGNAL-LEVEL REGULATION

The level of the signal at the input to the receiver in a PLC system will vary with attenuation on the line, and the amount of attenuation is not constant, but will vary with weather and other conditions, such as deposits on the insulators. Automatic gain-control (AGC) circuits are used to correct this problem.

In PLC systems, AGC is usually effected by comparing the level of a pilot signal against a fixed reference voltage at the receiver. In all cases, the pilot signal is transmitted at a level which is fixed with respect to the level of all other in-band signals, so the desired level is easily determined. The pilot signal used can be the pilot carrier used for frequency synchronization. In cases where the signaling tone is not interrupted, as when using frequency-shift keying, the signaling tone also may be used for AGC. RFL offers a choice of either system.

With the Model 65 AGC Automatic-Gain-Control Module a pilot tone is used to control the gain of a wideband amplifier through which the composite signal is passed.

Alternatively, the Model 65 FS SIG Frequency-Shift Signaling-Tone Transceiver can be equipped to provide AGC control by addition of an optional plug-on circuit card which contains an AGC circuit. The signal developed by this circuit is fed to the Model 65 Demodulator, where it holds the signal level constant by controlling the gain of an amplifier through which the composite signal passes.

With either system, the AGC control is effective over a range of 40 dB in signal-level variation. See Figure 5.1.

Chapter 2

THE TRANSMISSION MEDIUM

INTRODUCTION

The quality of a PLC system is determined by the noise and attenuation found in the transmission medium. And these also have a direct bearing on both the amount and kind of terminal equipment needed, and on the level of power transmitted for reliable communication. These are discussed in this chapter.

NOISE

Any long, unshielded transmission line will be subject to both inherent noise and atmospheric electrical noise. The various types of noise present on a powerline will contribute different types and degrees of degradation to the different communication functions of the PLC system. Some noise will be random and will appear as "white noise" within the carrier band. Such noise will produce hissing or rushing sounds in voice circuits, but it will not normally interfere severely with intelligibility. It will, however, contribute to the steady output level of detectors used in AM-type receivers, and it will affect S/N-ratio detectors. Random noise, caused by thermal agitation in the conductors, will add to similar noise peculiar to the electronic equipment used.

Other types of noise are specific to the powerline and system. There will be impulse noise due to electrical storms and to transients on the line caused by switches and circuit breakers. There will be corona noise, caused by ionization of the atmosphere and discharge across dirty or defective insulators. This will be worse on very-high-voltage lines. The severity of these types of noise will vary with weather conditions. There will also be some crosstalk noise and distortion products from harmonics and from intermodulation.

There is no reliable way to determine the noise level on a transmission line except to measure it, but a discussion of typical levels to be encountered and their likely effect on performance will be presented.

Impulse noise is more important in PLC applications than random noise. The noise impulses may occur at irregular intervals or may be periodic. Peak values of the impulses will typically be up to ten times the average noise level, and may present problems in triggering certain threshold devices, possibly contributing to false trips in line-protection equipment. In voice circuits they will appear as clicks or pops which do not destroy intelligibility. When impulses are very closely spaced in time, however, they tend to produce a steady crackling sound which masks speech, and the severity of masking is a function of both amplitude and frequency of the peaks. To measure the masking effect of such noise, meters have been devised which employ detectors with fast rise times and relatively slow decay times, so that the readings approach the peak amplitude for closely spaced impulses but fall off for more widely spaced pulses. Noise measured by such a meter is referred to as "quasi-peak" noise, and the time constants are so chosen that the measurement corresponds to the masking effect of noise on speech.

Irregularly spaced impulses may be caused by operation of switches or breakers, or they may be due to lightning strikes. Noise produced by the arc which follows the operation of a disconnect switch is particularly troublesome, with spikes of voltage up to 1000 volts, and with impulses repeated for up to a second in duration. Line faults may cause complex patterns of noise, with high-level impulses occuring at the time of fault and again at the operation of the breaker, while corona noise caused by ionization of the atmosphere fills the intervening period. Tuned circuits in the communication link will be excited to ringing by impulses, with narrowband channels tending to slower rise and decay times than broadband channels, but the latter show greater peak amplitude of oscillation.

Periodic impulse noise can be caused by rotating machinery, but it is usually associated with periodic variations in corona, or high-voltage discharge which follow the variations in peak voltage and hence occur at multiples of the powerline frequency. Such periodic corona not only causes noise itself, but also produces periodic fluctuations in the carrier-frequency impedance and in loss characteristics, with resultant power-frequency modulation of the carrier. The intermodulation products so produced are proportional to the carrier strength, so that it is not possible, for this type of noise, to improve the S/N ratio by increasing the signal strength.

A particularly troublesome type of noise has come to attention in recent years with the increased use of high-voltage dc power transmission (HVDC). Powerlines carrying dc, or, more commonly, ac powerlines which interconnect with or pass contiguous to such dc lines are subject to severe noise in the lower PLC spectrum, mainly below 100 kHz, arising from the converters or inverters employed. Such noise may reach levels of 20 to 30 dBm in the immediate vicinity of the converter station, and the only way a line so affected can be used practically for PLC in the lower frequency region is to isolate the sections with high noise by using expensive blocking filters constructed from tuned wave traps and coupling capacitors. Fortunately, the noise is fairly well localized near the converter stations, and conducted noise is primarily of the Mode-3 variety, described in the section following, titled Attenuation. Mode 3 fades rapidly at some distance from the source, so that more remote sections of the line may be suitable for normal PLC usage.

The power in white noise, corona noise, and most random-impulse noise is proportional to the bandwidth in which it is measured, so that the voltage amplitude is proportional to the square root of the bandwidth. Specification of noise is meaningless without specification of the bandwidth within which it is measured. Corona noise and, to some extent, other noise on a powerline tends to decrease with increasing frequency in the PLC band. This partially compensates for the increase in attenuation with increasing frequency.

Noise on powerlines propagates in the same manner as signal components at the same frequency, so that the natural modes of propagation discussed in connection with line attenuation are also applicable to noise attenuation. In estimating noise known to originate at a specific location, modal propagation should be taken into consideration.

Adverse weather will significantly increase the noise and attenuation of a powerline at carrier frequencies. Reliable operation under conditions of adverse weather will require providing for a large margin above the S/N ratio achieved in fair weather. It will normally be necessary to anticipate at least 20 dB higher noise levels in adverse weather. Noise produced by leakage across dusty insulators will typically increase during the first part of a light rainfall as the moisture increases the conductivity of the dust, but it will drop off somewhat as the insulators are washed clean. Humid weather without rainfall may cause severe leakage across dirty insulators, and this is especially true where certain types of chemical ash exist in the atmosphere. Lightning strikes may cause impulses 30 dB or more above the fair-weather peaks.

Noise may be measured on a functioning powerline by taking readings of radio noise (RN) with a field-strength instrument operating near the powerline. These data are converted, through available conversion tables or factors, to radio-interference voltage (RIV) on the line. Bandwidth corrections may have to be applied. Noise also may be measured on the communication-equipment side of the coupling capacitor, using an instrument such as the Electro-Metrics Model EMC-10, or the Singer Model NM-7 radio-interference meters.

In the absence of more specific data, the curves in Figure 2.1 may be used to estimate the probable worst-case, fair-weather noise for powerlines of various voltage ratings. They apply through the PLC frequency range. The noise powers shown in dBm are for a 3-kHz bandwidth. It is estimated that noise will fall below the indicated values 75% of the time. For adverse weather, at least 20 dB should be added to the fair-weather values.

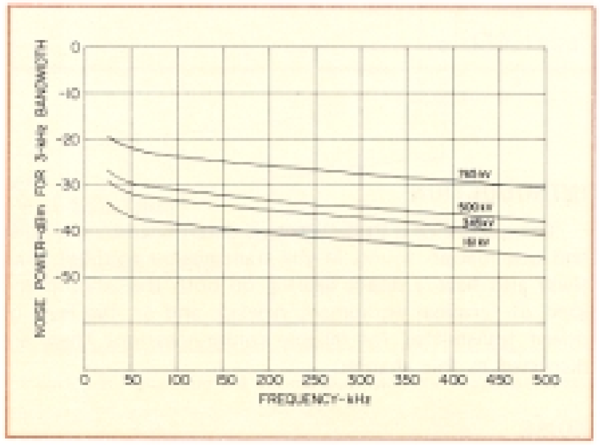

Figure 2.1. Probable worse-case, fair-weather noise for powerlines of various voltages.

ATTENUATION

Until the early 1960's, the subject of carrier-frequency attenuation on powerlines was not well understood, and attempts to apply standard transmission-line theory had not yielded results in good agreement with measured values. It was then common practice to predict line attenuation on the basis of charts and tables, using correction factors derived from averages of many experimentally determined parameters on lines of various types.

Average values of parameters which exhibit a rather wide and unexplained variability are not very useful to an engineer during the planning stages of a new system, and it was fortunate that a more adequate theory was developed, in the early 1960's, by applying the methods of matrix analysis. This resulted in the modal theory of signal propagation. Many papers have been published on the theory of modal analysis, and good correlation of theory and measured parameters has been achieved in most cases.

Much of the published information on modal theory has been the work of engineering scientists in the research laboratories of large corporations or within the engineering departments of educational institutions, and the sophistication of the terminology and the mathematical methods employed may have made the work seem formidable, if not inaccessible, to those who have not specialized in these fields. On the other hand, papers which have attempted to present a more generalized and simplified approach may not have provided sufficiently specific information to be of genuine assistance to the engineer who wishes to apply the theory to his practical problem.

In this guide, the subject of line attenuation will be approached from the viewpoint of modal theory, but charts and tables and straightforward formulas, together with worked out examples, will be presented to make the subject easily comprehensible at the same time that it is specifically useful.

To estimate the carrier-frequency attenuation of a multiconductor system, such as a single- or double-circuit powerline, it is necessary to develop a system of analysis which takes into account the mutual coupling between the conductors.

The signal at any point on any conductor of a system of N signal-carrying conductors consists of N superimposed components, called "modal components," each of which may propagate with its own characteristic attenuation per unit length and its own characteristic velocity. The situation is somewhat analogous to the propagation of a complex waveform in a non-ideal transmission medium, in which analysis of the propagation requires reduction of the waveform into its frequency components by Fourier analysis so that the attenuation and time delay of each of the frequency components may be separately considered, and the final waveform may be established by recombination in the time domain. Each signal component on each conductor of a multiconductor system results from the self and mutual impedances associated with that conductor, and propagates in a mode which is natural to the system and which is determined by physical parameters of the system, such as conductor resistance and geometry, system geometry, resistivity of the underlying earth, and others.

If the voltages on all N conductors of a system at some point are known, they can be resolved into their modal components by making use of the property of a pure mode of propagation that the ratio of the voltages on any pair of conductors is everywhere constant and fixed for that mode. If the system is energized so that the initial conductor voltages exhibit the ratios of a pure natural mode of propagation for that system, then only this single mode will exist and propagate unless a discontinuity occurs which generates other modes. The excitation produced by most conventional coupling arrangements results in the production of more than one mode of propagation.

The voltage (or current) ratios and the complex propagation constants characteristic of each of the N modes can be computed mathematically by expressing the N propagation equations in matrix form and solving for the eigenvectors, which establish the voltage ratios, and their associated eigenvalues, which are the complex propagation constants. Calculation of these eigenvectors and eigenvalues involves very laborious mathematics, and a digital computer is definitely required.

A method of calculation described by Galloway, Shorrocks, and Wedepohl (1) has been used in the power-communications industry with generally good agreement between calculated and measured modal parameters. RFL has implemented this method on a digital computer and used it to predict the modal parameters of a number of different lines, both as part of the Company's research program and in response to tenders from the utilities involved. Plots of the three most critical modal parameters, generated from this computer analysis on five different powerlines, are presented in this section along with some discussion to serve as a guide to engineers planning a new installation or desiring to analyze an existing one. When these three parameters have been estimated, charts and simple formulas will enable an engineer with a scientific calculator to estimate with good accuracy the attenuations and optimum coupling configurations for any three-phase horizontal powerline typically encountered. For special, more complex cases, RFL can supply more rigorous computer calculations.

(1) R. H. Galloway, W. B. Shorrocks, and L. M. Wedepohl, "Calculation of Electrical Parameters for Short and Long Polyphase Transmission Lines." Proc. IEE (London), Vol. 111, No. 12, pp. 2051-2059, (Dec. 1964).

Several considerations will emphasize the significance of modal analysis in estimating PLC attenuation. First, the various modes will propagate down the line with different efficiencies. The distribution of the applied signal power among the various modes will be a function of the phase or phases chosen for coupling, and it is evident that coupling should be accomplished in such a way as to distribute a maximum part of the applied power into the most efficient mode or modes.

Another consideration arises because the modes propagate with different velocities, so that there are phase shifts which result in reinforcements and cancellations between the modal components of a signal. These may cause fluctuations in attenuation at different frequencies and at different locations, which appear like standing waves, even though the line may be correctly terminated. The signal minima associated with such cancellations are not actually power losses, but transfers of signal power from one phase to another.

Transpositions in the line produce modal conversions because of the redistribution of signal among the phases. Modal components will frequently cancel partially on one phase at the same point that they reinforce on another phase, so that the choice of phase becomes very significant. Quite often, the greatest available signal at a receiver will be on a phase or pair of phases other than those chosen for coupling at the transmitter. No discussion of powerline carrier attenuation can be adequate if it is based on averages of measured values without taking into account the modal nature of the signals.

To demonstrate the essential nature of modal analysis, we will consider the most common case for powerline carrier of a single-circuit, horizontal, three-phase powerline. This case has been discussed extensively in the literature. The natural modes of propagation for such a system are as follows:

Mode 1

This mode consists of signal power which propagates out equally on the two outer phases and returns on the center phase, or vice versa. This is an approximation, but it is a valid one for this type of line. Mode 1 is the most efficient mode, and it propagates at essentially the speed of light. Mode-1 coupling efficiency is a measure of the proportion of incident power distributed to Mode 1, and we normally wish to keep this as high as possible. For long, untransposed lines, only Mode-1 power will be available at the receiver. In some published papers, this mode is referred to as Mode 3.

Mode 2

This mode consists of signal which propagates out over one outside phase wire and returns over the other, with no signal for Mode 2 on the center conductor. The propagation efficiency and velocity for Mode 2 are between the values for Mode 1 and Mode 3. On long, untransposed lines most of the Mode-2 power is lost.

Mode 3

Mode 3 is sometimes referred to as the ground mode. The signal propagates out essentially equally on all three phases and returns via the ground. Because of the high loss of the ground, this signal component is severely attenuated, and Mode 3 is essentially useless. We must take this mode into consideration at the points of coupling and transposition, however, in order to determine how much of the signal power is lost to this mode. This loss will be regarded as part of the insertion loss of the system. In some papers, this mode is referred to as Mode 1.

The diagram in Figure 2.2 summarizes the modal components of the three modes for a single-circuit, horizontal, three-phase powerline. The component voltages are normalized to the voltage on Phase 1.

For most systems, there will be significant quantities of Mode-1 and Mode-2 voltage at the terminus, or at the transposition. These voltage components will add on one of

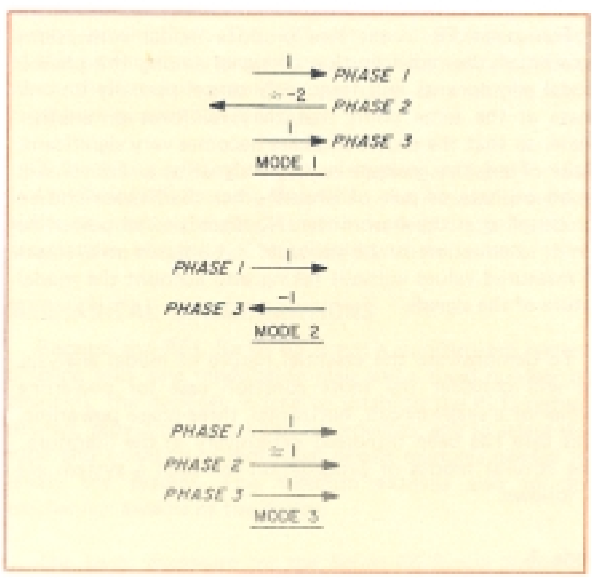

Figure 2.2. Modal components of the three modes for a single-circuit, horizontal, three-phase powerline.

the outer phases and subtract on the other. The voltages on the outer phases will be vector sums of the modal components, and their magnitude will depend on the coupling as well as on the phase shift and attenuation. These in turn, will depend on the frequency and the length of the line section. A few "rules of thumb" become immediately evident.

For single-phase to ground coupling, it is almost always most efficient to couple to the center phase, since this will maximize the power coupled into Mode 1 while producing no power in Mode 2. A "push-push" coupling arrangement to the outer phases is also efficient, but it requires more coupling equipment. For phase-to-phase coupling (push-pull), it is normally most efficient to couple between the center phase and one outer phase, since coupling to the two outer phases produces pure Mode 2. This is more efficient than coupling center-phase to ground, but again it requires more coupling equipment. Sometimes a need for reliability will dictate phase-to-phase coupling so that not all of the signal power is lost in case of a single line-to-ground fault near the transmitter or the receiver. This is particularly important if the PLC system is used for line protection.

Choices of frequency and of coupling phase will be facilitated by knowledge of the frequencies of modal reinforcement or cancellation between Modes 1 and 2. The frequencies depend on the relative velocity, K, of Mode 2 with respect to Mode 1. The factor K can be estimated with the aid of the typical line-parameter examples to be given in this chapter. Then, the critical frequency for a line, or section, of length L may be found from:

$$F = \frac{Kc}{2L\ (1-K)} \tag{2.1}$$

where c is the velocity of light. The frequency, F, and its odd multiples are frequencies at which the modal-components will partially cancel on the outside phase to which coupling was made, while the components will be reinforced on the opposite outside phase. For even multiples of F, the opposite will be true.

In order to assure a maximum proportion of the most efficient mode of propagation, it is always desirable to have as high a signal strength as possible on the center-phase conductor. At a transposition, one of the outside phases will be transposed to center, and it is preferable that this be the outside phase with the higher signal voltage. This principle, together with Equation (2.1), may assist in choosing the optimum outside phase for phase-to-phase coupling.

Derivation of Equations for Horizontal Three-Phase Line

If we use the voltage ratios shown in Figure 2.2, and establish Phase 1 as the reference phase, we may express the voltages on the three phases by the following three equations

$$P_1 = M_1 + M_2 + M_3 \quad (2.2)$$

$$P_2 = M_3 - 2M_1 \quad (2.3)$$

$$P_3 = M_1 - M_2 + M_3 \quad (2.4)$$

where $M_{(i)}$ is the modal component of voltage on Phase 1 for Mode (i). Solving for the "modal voltages":

$$M_1 = (P_1 - 2P_2 + P_3)/6 \quad (2.5)$$

$$M_2 = (P_1 - P_3)/2 \quad (2.6)$$

$$M_3 = (P_1 + P_2 + P_3)/3 \quad (2.7)$$

For simplified calculations, it is safe to ignore the M_3 terms, since Mode 3 is severely attenuated and no Mode-3 voltage normally reaches the end of the line section.

Estimating the Modal Parameters

Before it is possible to calculate the attenuation of a powerline on a modal basis, it is necessary to estimate the most important modal parameters. These are the modal attenuation constants, in dB per kilometer, for Modes 1 and 2, and the modal velocity ratio, K. The plots of five typical lines, given in Figures 2.3 through 2.8, will assist in this estimation. For each of the five lines presented, two key line parameters are shown. These are the distance-to-average-height ratios of the phase conductors, and the conductor-resistance factor, R_F. The average height is generally computed as the maximum height (at the tower) minus two-thirds of the average mid-span sag. The resistance factor, R_F, multiplied by the square root of the frequency gives the resistance coefficient, in ohms per meter, of the phase conductors (1). More complete information on the lines can be found in Table 2.1.

Figures 2.3 through 2.8 will show, in general, how the modal parameters are affected by variations in earth resistivity. Mode-1 attenuation and relative Mode-2 velocity increase with decreasing earth resistivity. For Mode-2 attenuation, the situation is more complex, and there is normally a value of earth resistivity, for any given configuration of line, at which the earth's contribution to Mode-2 losses will be at a maximum.

Line A is a 400-kV line of twin, bundled ACSR conductors over varied terrain, and the modal parameters were calculated for various earth resistivities as measured by the utility.

Lines B and C are 500-kV lines with bundled ACSR conductors. These two lines have very similar resistance factors, somewhat lower than Line A, and their modal attenuations are generally lower and their relative velocities higher. Line C has a low d/h ratio, which may have contributed to its low modal losses and high K factor.

TABLE 2.1

PHYSICAL CHARACTERISTICS OF SOME TYPICAL HIGH-VOLTAGE TRANSMISSION LINES

Line	A	B	C	D	E_1	E_2
Voltage – kV	400	500	500	220	115	115
Conductors/Phase	2	4	3	1	1	1
Conductor Size – MCM	1034	500	954	795	4/0 AWG	267
Radius of Bundle – mm	226	323	188	–	–	–
R Factor, R_F, Ohms/Meter x 10^{-6}	2.38	1.94	1.94	6.38	13.36	11.01
Max. Height of Conductors – M	28.6	29.8	–	16.7	–	–
Sag of Conductors – M	11.7	18.6	–	11.8	–	–
Avg. Ht. of Conductors – M	12.8	16.8	18.1	9.36	9.14	9.14
Spacing of Conductors – M	11.5	12.2	9.75	7.11	3.81	3.81
Ratio d/h	0.902	0.736	0.538	0.824	0.417	0.417
Size of Gnd. Wires	11 mm	5/16" (7.93 mm)	3/8" (9.53 mm)	3/8" (9.53 mm)	3/8" (9.53 mm)	3/8" (9.53 mm)
Max. Height Gnd. Wires – M	29.1	39.3	–	24.9	–	–
Sag of Gnd. Wires – M	9.0	15.5	–	9.75	–	–
Avg. Height Gnd. Wires – M	23.1	29.8	29.9	18.4	12.2	12.2
Gnd Wire Spacing from Center – M	±7.93	±9.53	±10.7	±3.25	±2.06	±2.06

Line D is a 220-kV line with typically much higher resistance factor than the EHV lines. Conductor resistance is an important constituent of Mode-1 attenuation, and the lower voltage lines show higher Mode-1 losses, as would be expected. The d/h ratio of Line D is similar to Lines A and B, and the relative velocity is also in the same region. Mode-2 attenuation of the 220-kV Line D seems to be affected somewhat differently by earth resistivity, from the case with the EHV lines.

Line E is a 115-kV line suspended from wood H towers with relatively higher resistance factor and lower d/h ratio. Two sizes of conductors were employed, designated E_1 and E_2, and the Mode-1 attenuation reflects the difference in resistance. The relative Mode-2 velocity is high, probably because of the low d/h ratio. The Mode-2 attenuation is low, is little affected by resistance, and is similar to Line C which also had a low d/h ratio.

As more lines are analyzed, it is planned that more complete sets of curves may be made available, and that more specific procedures for interpreting and interpolating the results may be offered. Meantime, the results presented here should enable the engineer to make a more reasonable estimate of the probable modal constants of his powerline than is possible from a mere presentation of the likely range of the maximum variation of each parameter.

(1) For ACSR conductors, the resistance factor, R_f, may be calculated from the following:

$$R_f = 2.55 \times 10^{-7} / (n_b \times r_s \times (n_s + 2)),$$

when n_b is the number of subconductors per phase, for bundled conductors, r_s is the radius in meters of each strand of the conductor, and n_s is the number of conductor strands in the outer layer.

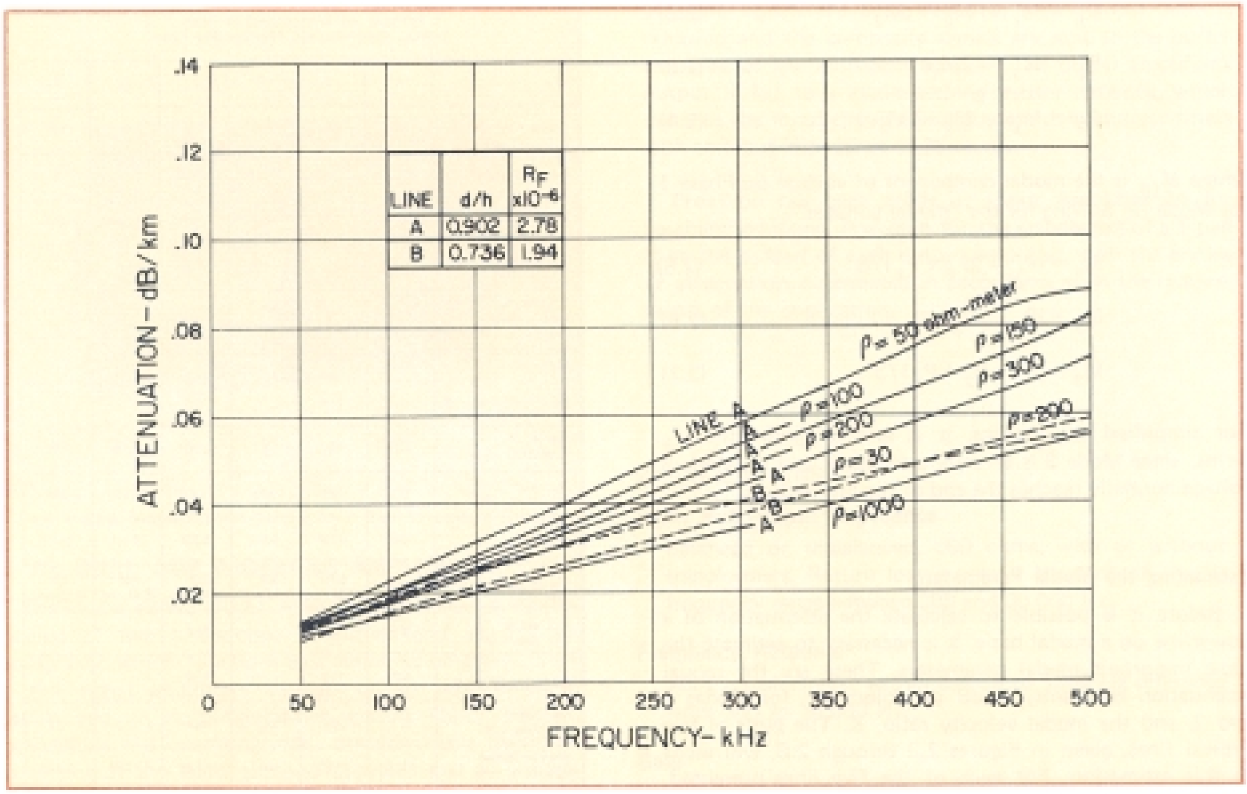

Figure 2.3. Mode-1 attenuation, α_1, for Lines A and B.

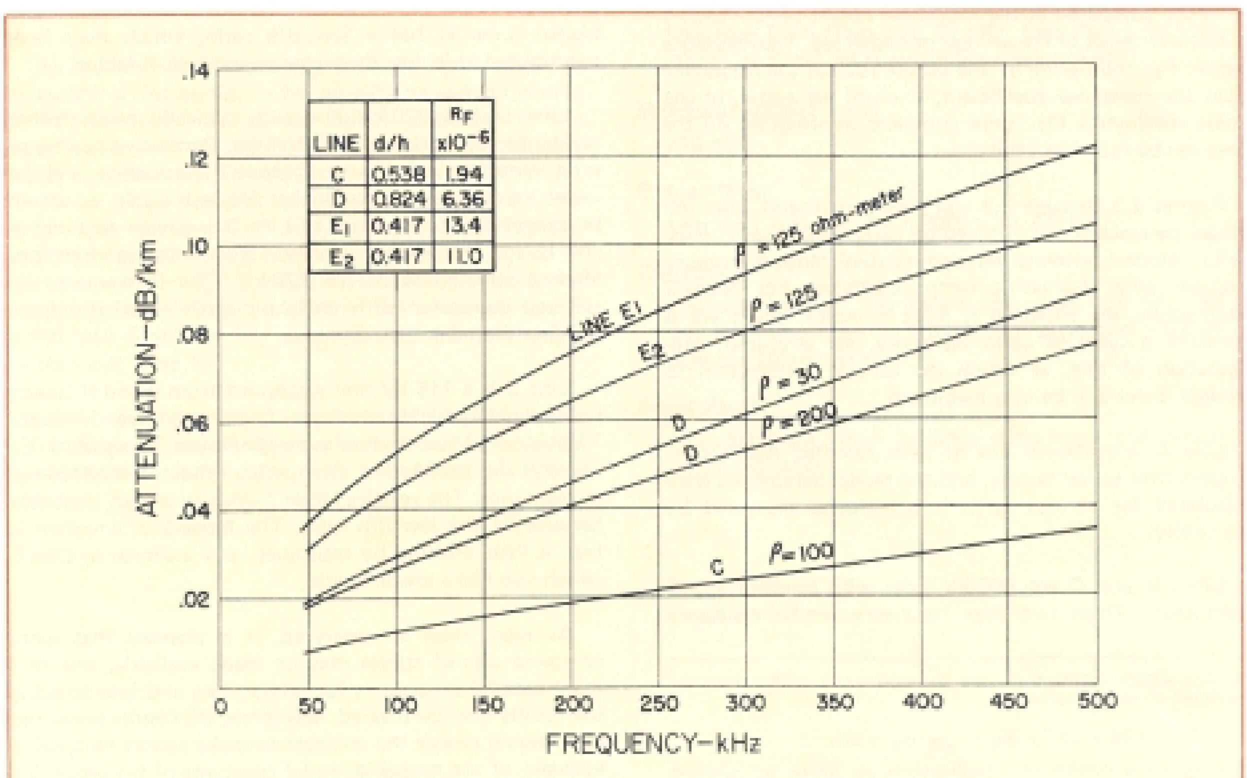

Figure 2.4. Mode-1 attenuation, α_1, for Lines C, D, E_1, and E_2.

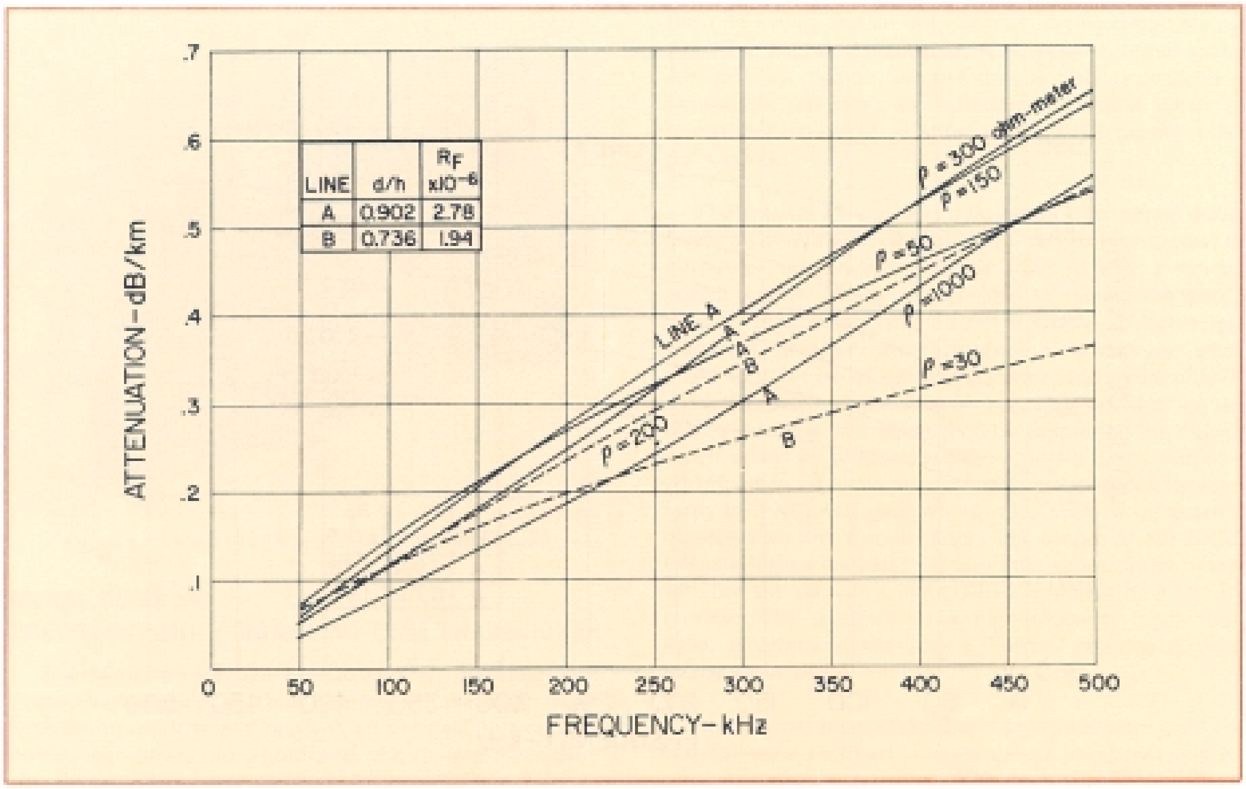

Figure 2.5. Mode-2 attenuation, α_2, for Lines A and B.

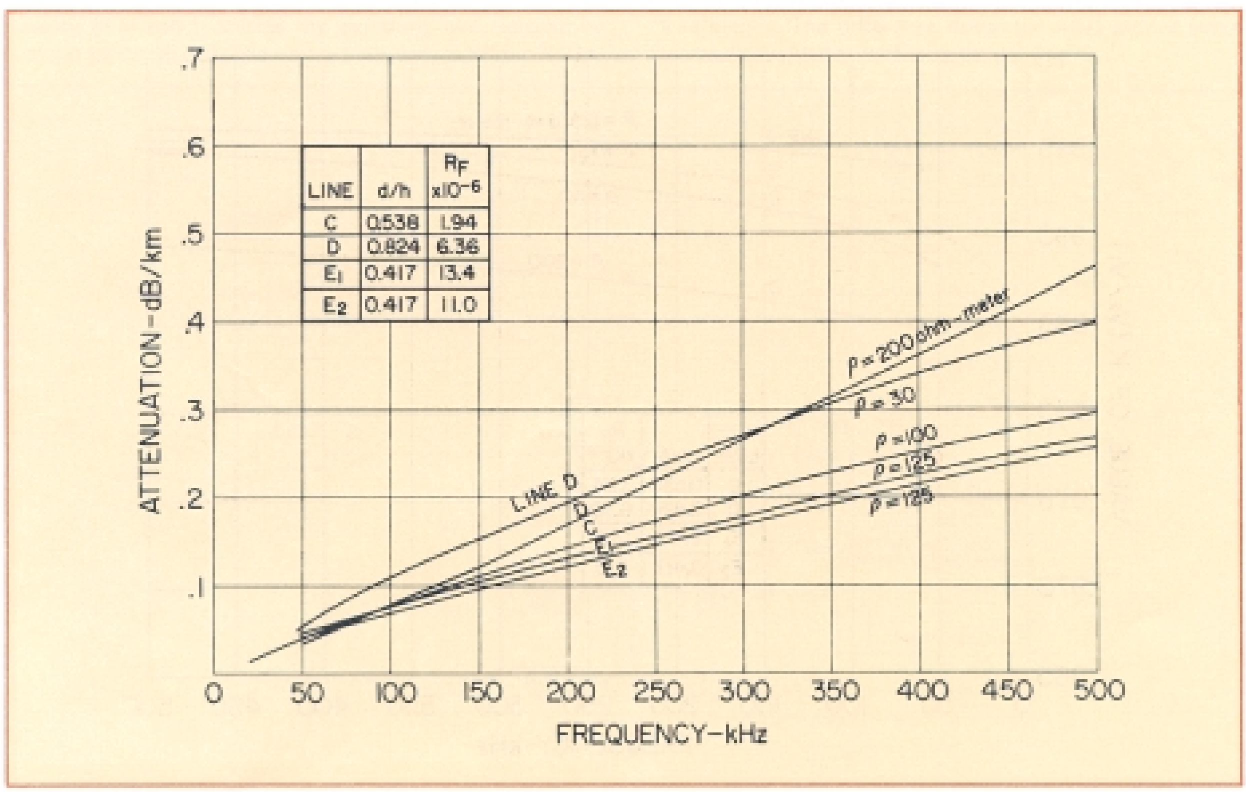

Figure 2.6. Mode-2 attenuation, α_2, for Lines C, D, E_1, and E_2.

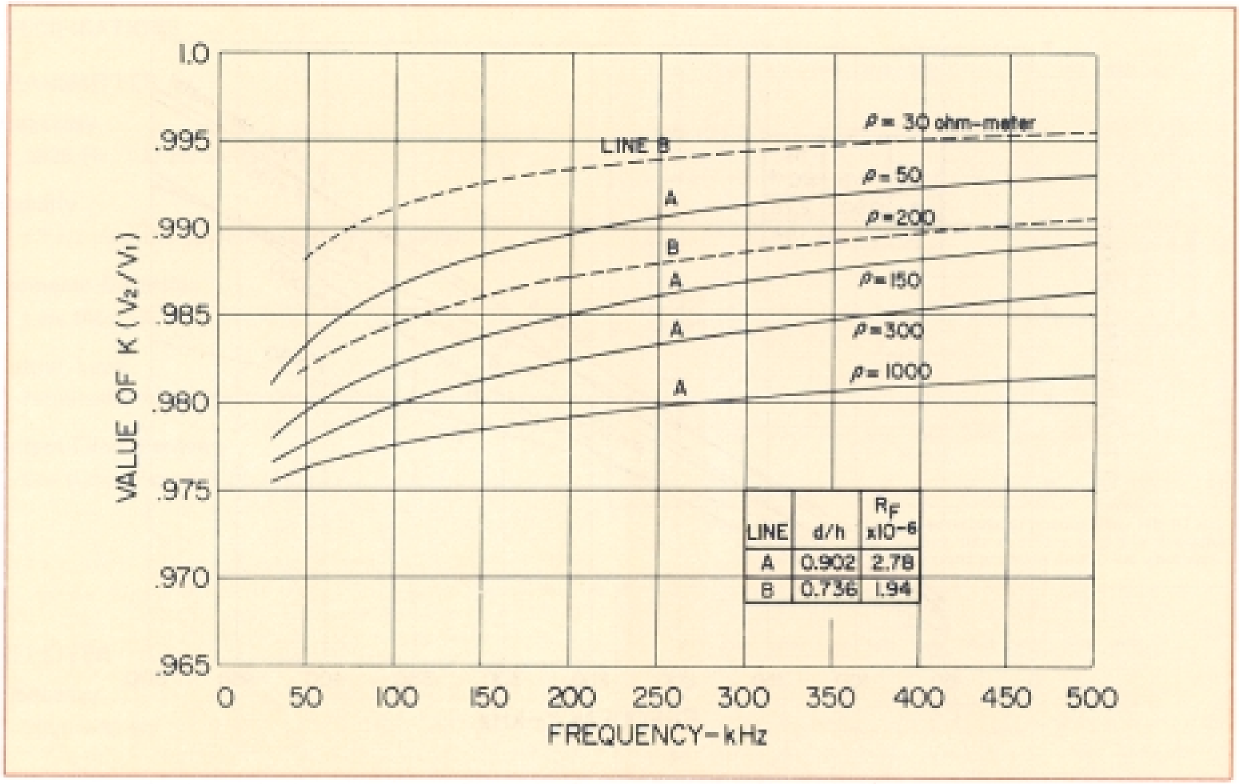

Figure 2.7. Relative velocity, K, of Mode 2 with respect to Mode 1 for Lines A and B.

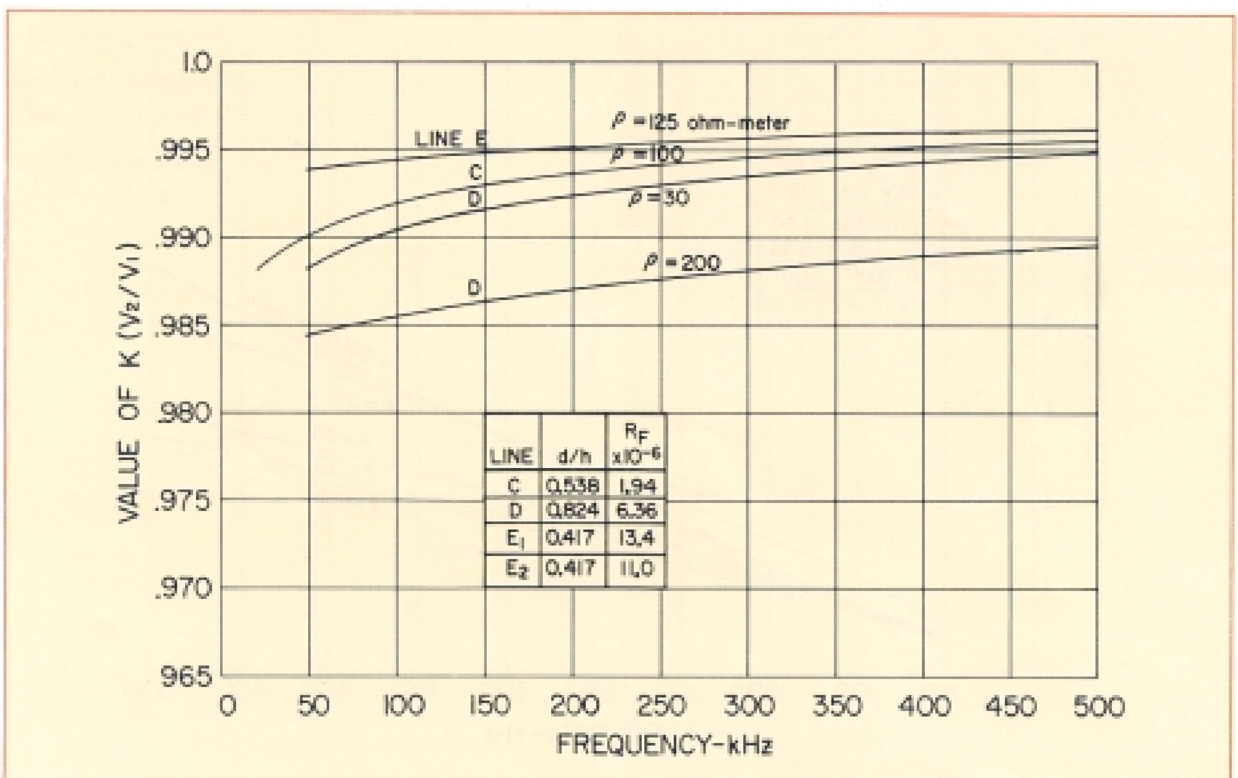

Figure 2.8. Relative velocity, K, of Mode 2 with respect to Mode 1 for Lines C, D, and E.

Calculating Attenuation

Calculation of line attenuation on a modal basis may be summarized briefly: The phase voltages at the point of excitation, which are usually normalized to the voltage on Phase 1, are converted to modal voltages. The attenuation and phase shift of each modal voltage at the end of the section is calculated, and then the modal voltages are converted back into the equivalent phase voltages, where they are transposed, if required, and then used either as input for another section or to compute the attenuation. Since the conversions between phase and modal voltages involve the addition and subtraction of complex voltages, it is mathematically convenient to express them in rectangular form. Since the computation of the modal losses involves the multiplication of complex voltages by complex propagation coefficients, it is more convenient to express them here in polar form. Accordingly, the phase-to-modal conversions, and vice versa, are interspersed with rectangular-to-polar conversions, and vice versa.

RFL Form PLC-128A has been prepared to assist an engineer planning a three-phase PLC system to estimate line attenuation on a modal basis. Copies of this form may be obtained from RFL, and a sample has been worked out for a two-section, singly transposed, horizontal, three-phase line. This form, and the equations given thereon, can be used with any scientific or engineering calculator to give a very good approximation of line attenuation, even though some simplifying generalities have been employed. Radian measure of angles is used here, and care must be taken that the signs of the rectangular coordinates and the quadrants of the angles are correct.

For the line analyzed as a sample in Figures 2.10 and 2.11, in which each line section is 80 miles long and the analysis is at 80 kHz, the modal attenuation constant of Mode 1 was estimated, with the aid of Figure 2.3, at 0.0174 dB/km. The constant for Mode 2 was estimated, from Figure 2.5, at 0.0808 dB/km, and the ratio K was estimated from Figure 2.7 at 0.985. The phase voltages at the input are normalized to the voltage on Phase 1, and push-pull, phase-to-phase coupling is applied to Phases 1 and 2. Input voltages are considered real.

The first step in the use of the form is to convert to modal voltages for Modes 1 and 2, using Equations (1) and (2), shown on the form. The equations must always be applied to both the real and the imaginary parts of the phase voltages. Mode 3 is ignored, since it will be completely attenuated. The real and imaginary parts of Mode-1 voltage are entered in Boxes 7 and 8, and the Mode-2 voltages are entered in Boxes 9 and 10. Equations (13) and (14) may be used here to convert the rectangular modal voltages to polar form, and the polar values entered in Boxes 11 and 12 for Mode 1 and Boxes 13 and 14 for Mode 2.

The attenuations for Modes 1 and 2, based on the section length, are found from the estimated attenuation constants, using Equations (6) and (7), and may be entered at those points. The unit of length employed is kilometers, to agree with international usage. The voltage ratios corresponding to these attenuations may be calculated from Equation (10) and entered in Boxes 15 and 16. The modal voltage magnitudes at the end of the section are found by multiplying the values in Boxes 11 and 13 by the ratios in Boxes 15 and 16, and are entered in Boxes 18 and 19.

The estimated value of K, together with the length and frequency, is used in Equation (8) to compute the phase shift of Mode 2 with respect to Mode 1, and this is entered in Box 17. Relative phase shift is used to simplify the calculations, since we are not concerned with the absolute phase shift of the system. The relative phase angle at the end of the section is determined from the shift and from the initial angles in Boxes 12 and 14, using Equation (9). This is entered in Box 20. Since we have normalized the phase differential with respect to Mode 1, we can transfer the value from Box 18 to Box 21. The rectangular values for Mode 2, in Boxes 22 and 23, are determined by polar-to-rectangular conversion of the values in Boxes 19 and 20, Equations (11) and (12).

The final step for this first line section is to convert the two modal voltages in Boxes 21, 22, and 23 to the three phase voltages at the end of the section. Equations (3), (4), and (5) are used for this conversion, applied to both the real and imaginary parts, and the values are entered in Boxes 24 through 28. The center-phase voltage is real because we have normalized the phase shift with respect to Mode-1, and only Mode-1 voltage appears on the center conductor. Since there is no imaginary component to Mode 1, the values in Boxes 27 and 28 are the positive and negative of the value in Box 23.

The complex phase voltages at the end of the section, Boxes 24-28, can now be transposed in accordance with the type of transposition employed, and they are entered into Boxes 1 through 6 of the second sheet, Figure 2.11, which is used for the second section of the line. The values in Boxes 15, 16, and 17 can be transferred from Sheet 1, since the second section is the same length as the first. Sheet 2 is filled in exactly as Sheet 1, but, at the end, the complex phase-to-phase voltages can be computed and entered in Boxes 35 through 40. The magnitudes of the three phase-to-phase voltages can be computed from Equation (13) and entered in Boxes 41, 42, and 43.

The attenuations for the phase-to-phase couplings are computed with reference to the input phase-to-phase voltage, which was two volts, and entered in Boxes 44, 45, and 46. The designations of the phase conductors as 1, 2, and 3 refer here to the spatial position of the phases, and they do not imply physical continuity of the phase wires. The phase-to-phase combination 2-3, at the end of this line, is the same physical pair to which we coupled at the transmitting end. It will be noted that the attenuation on this pair is much higher than the attenuation on Phase Pairs 1-2 or 1-3. Figure 2.9 shows frequency response of this line from 30 to 230 kHz, showing the attenuations on the three phase-to-phase combinations at the output. The frequency of modal cancellation, computed from Equation (2.1), is 76.4 kHz. The null which occurs at this frequency on Phase Pair B-C (2-3) is obvious in Figure 2.9.

Although modal analysis will provide a much more accurate prediction of PLC-line attenuation than older methods based only on empirical data, it will depend for its accuracy on a knowledge of certain key parameters of the line. Of these, the most sensitive is the relative modal velocity. This parameter, in turn, is dependent upon such factors as line height and ground resistivity, and these may vary with line load, temperature, precipitation, etc., so that it will always be wise to be conservative in applying the results of analysis to prediction of worst-case losses.

Two methods now are available to those desiring to calculate the attenuation to be expected on a proposed or existing transmission line from a knowledge of the physical characteristics of the line and its environment. These are:

(a) Using the curves and table given in the foregoing discussion and Form PLC-128A, available from RFL, the attenuation may be easily calculated with the aid of a simple scientific computer. A reproducible copy of Form PLC-128A is included. Additional copies are available.

(b) A computer program has been developed by RFL to perform such calculations of modal attenuation in order to provide complete frequency-response data of the PLC line for various coupling arrangements. RFL is in a position to perform these calculations providing the information needed to characterize the line is given. Needed data are:

(a) Number of phase wires and earth wires.

(b) Vertical and horizontal coordinates of phase and earth wires at the tower.

(c) Average midspan sag of phase and earth wires.

(d) Number of subconductors per bundle.

(e) Radius of subconductor, of bundle, and of outer strand of subconductor.

(f) Number of strands in outer layer of subconductor.

(g) Radius of earthwire, of outer strand of earthwire, and the number of outer strands.

(h) Resistivity of underlying earth.

(i) Frequencies of interest.

(For standard conductors, information for Items (e), (f), and (g) can be obtained from standard wire tables.)

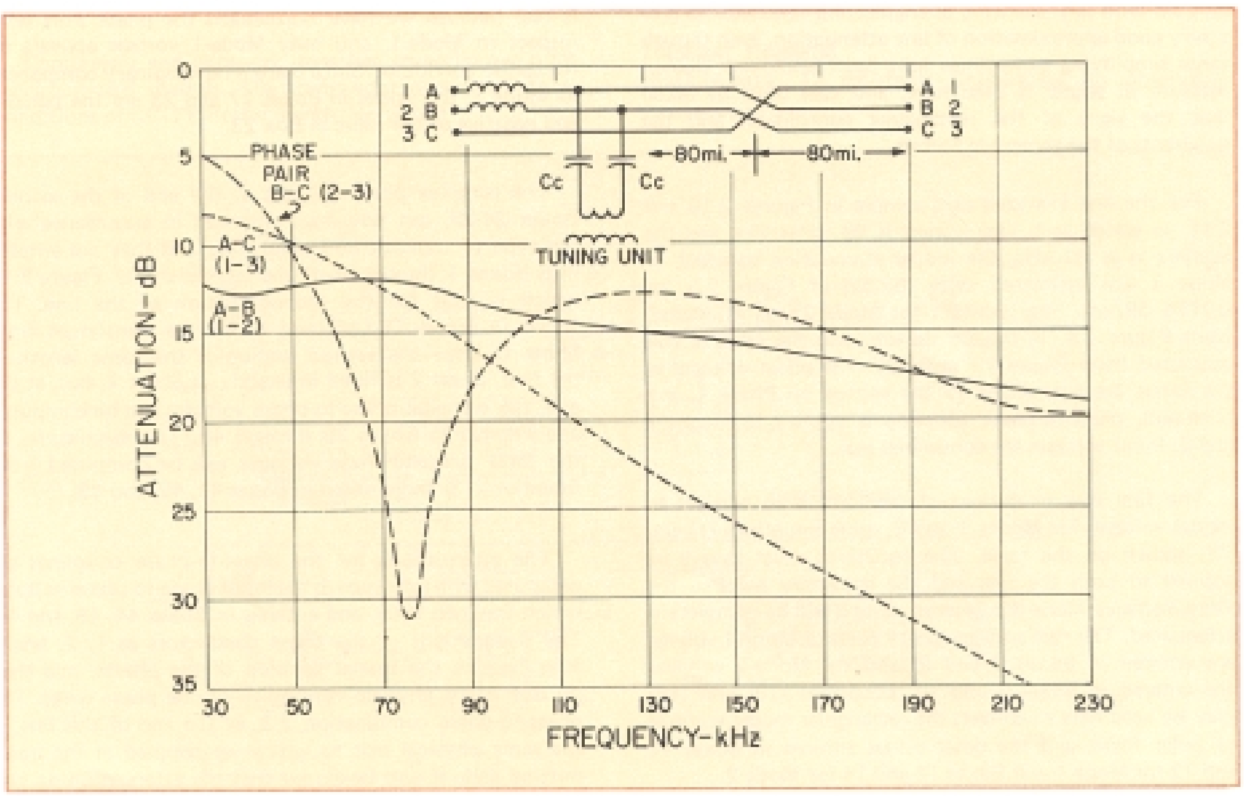

Figure 2.9. Frequency response of powerline used as an example.

MODE OR PHASE	START OF SECTION: PHASE VOLTAGE, P — REAL	PHASE VOLTAGE, P — IMAG	MODAL VOLTAGE, M — REAL	MODAL VOLTAGE, M — IMAG	MODAL VOLTAGE, M — MAG	MODAL VOLTAGE, M — $\angle\theta$	TRANSFER FUNCTION: ATT VOLT RATIO	TRANSFER FUNCTION: SHIFT	END OF SECTION: MODAL VOLTAGE, M — MAG	MODAL VOLTAGE, M — $\angle$	MODAL VOLTAGE, M — REAL	MODAL VOLTAGE, M — IMAG	PHASE VOLTAGE, P — REAL	PHASE VOLTAGE, P — IMAG	PHASE VOLTAGE, P — MAG	PHASE VOLTAGE, P — ATT
1	[1] 1.0	[4] 0.0	[7] 0.5	[8] 0.0	[11] 0.5	[12] 0.0	[15] 0.773		[18] 0.386	→	[21] 0.386		[24] 0.237	[27] $\overline{0.0228}$	[29]	[30]
2	[2] −1.0	[5] 0.0	[9] 0.5	[10] 0.0	[13] 0.5	[14] 0.0	[16] 0.302	[17] 3.29	[19] 0.151	[20] 3.29	[22] −0.149	[23] $\overline{0.0228}$	[25] −0.773		[31]	[32]
3	[3] 0.0	[6] 0.0											[26] 0.536	[28] 0.0228	[32]	[34]

PHASE PAIRS	REAL	IMAG	MAG	ATT
1-2	[35]	[38]	[41]	[44]
2-3	[36]	[39]	[42]	[45]
1-3	[37]	[40]	[43]	[46]

EQUATIONS

1. $M_1 = (P_1 - 2P_2 + P_3)/6$
2. $M_2 = (P_1 - P_3)/2$
3. $P_1 = M_1 + M_2$
4. $P_2 = -2M_1$
5. $P_3 = M_1 - M_2$
6. $ATT_1 = L\alpha_1 = \underline{2.24}$ dB
7. $ATT_2 = L\alpha_2 = \underline{10.4}$ dB
8. $SHIFT = 0.021\left(\frac{1-K}{K}\right)LF$
   WHERE L = LENGTH, km
   F = FREQ, kHz
9. $ANGLE = \theta_2 - \theta_1 + SHIFT$, RADIANS
10. $VOLTAGE\ RATIO = 10^{-\left(\frac{ATT}{20}\right)}$
11. REAL = MAG x cos(angle)
12. IMAG = MAG x sin(angle)
13. $MAG = \sqrt{(REAL)^2 + (IMAG)^2}$
14. $ANGLE = TAN^{-1}(IMAG/REAL)$

LENGTH 80 mi, 128.72 km
FREQ 80 kHz
α_1 = 0.0174 dB/km
α_2 = 0.0808 dB/km
K = 0.985

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ESTIMATE OF LINE ATTENUATION
SECTION 1
FORM PLC-128A
SHEET 1 OF 2

Figure 2.10. Calculation for Section 1 of powerline used as example.

MODE OR PHASE	START OF SECTION: PHASE VOLTAGE, P — REAL	PHASE VOLTAGE, P — IMAG	MODAL VOLTAGE, M — REAL	MODAL VOLTAGE, M — IMAG	MODAL VOLTAGE, M — MAG	MODAL VOLTAGE, M — $\angle\theta$	TRANSFER FUNCTION: ATT VOLT RATIO	TRANSFER FUNCTION: SHIFT	END OF SECTION: MODAL VOLTAGE, M — MAG	MODAL VOLTAGE, M — $\angle$	MODAL VOLTAGE, M — REAL	MODAL VOLTAGE, M — IMAG	PHASE VOLTAGE, P — REAL	PHASE VOLTAGE, P — IMAG	PHASE VOLTAGE, P — MAG	PHASE VOLTAGE, P — ATT
1	[1] 0.536	[4] 0.0228	[7] −0.119	[8] 0.0114	[11] 0.119	[12] 3.046	[15] 0.773		[18] 0.0921	→	[21] 0.0921		[24] 0.283	[27] 0.0516	[29]	[30]
2	[2] 0.237	[5] $\overline{0.0228}$	[9] 0.654	[10] 0.0114	[13] 0.654	[14] 0.0174	[16] 0.302	[17] 3.29	[19] 0.198	[20] 0.264	[22] 0.191	[23] 0.0516	[25] −0.184		[31]	[32]
3	[3] −0.773	[6] 0.0											[26] $\overline{0.0987}$	[28] 0.0516	[32]	[34]

PHASE PAIRS	REAL	IMAG	MAG	ATT
1-2	[35] 0.467	[38] 0.0516	[41] 0.470	[44] 12.6
2-3	[36] $\overline{0.0855}$	[39] 0.0516	[42] 0.0999	[45] 26.0
1-3	[37] 0.382	[40] 0.103	[43] 0.395	[46] 14.1

EQUATIONS

1. $M_1 = (P_1 - 2P_2 + P_3)/6$
2. $M_2 = (P_1 - P_3)/2$
3. $P_1 = M_1 + M_2$
4. $P_2 = -2M_1$
5. $P_3 = M_1 - M_2$
6. $ATT_1 = L\alpha_1 = \underline{2.24}$ dB
7. $ATT_2 = L\alpha_2 = \underline{10.4}$ dB
8. $SHIFT = 0.021\left(\frac{1-K}{K}\right)LF$
   WHERE L = LENGTH, km
   F = FREQ, kHz
9. $ANGLE = \theta_2 - \theta_1 + SHIFT$, RADIANS
10. $VOLTAGE\ RATIO = 10^{-\left(\frac{ATT}{20}\right)}$
11. REAL = MAG x cos(angle)
12. IMAG = MAG x sin(angle)
13. $MAG = \sqrt{(REAL)^2 + (IMAG)^2}$
14. $ANGLE = TAN^{-1}(IMAG/REAL)$

LENGTH 80 mi, 128.72 km
FREQ 80 kHz
α_1 = 0.0174 dB/km
α_2 = 0.0808 dB/km
K = 0.985

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ESTIMATE OF LINE ATTENUATION
SECTION 2
FORM PLC-128A
SHEET 2 OF 2

Figure 2.11 Calculation for Section 2 of powerline used as example.

OTHER LOSSES

In addition to attenuation, it is necessary to take into account other losses within the system.

The entrance cable may contribute losses ranging from 0.3 to 1.6 dB per 1000 feet. Resistive losses in the coupling capacitors and in the line-tuning units may be conservatively estimated at 3 dB per coupling, as discussed in the following section titled Coupling. Coupling losses at the receiving end need not be considered in making estimates of signal-to-noise ratio, because noise and signal will be equally attenuated. Additional shunt losses may contribute 1 to 3 dB, depending on the efficiency of the line traps used. Any line taps employed should be completely bypassed by the use of additional traps and coupling capacitors, as they may otherwise act as resonant stubs and contribute unpredictable variations in impedance and standing waves. The losses of such bypasses may be 6 dB or more per incident, and if this is unacceptable then some form of repeater should be employed.

Adverse weather, in addition to increasing noise, will tend to increase attenuation at PLC frequencies. Damp weather and early light rainfall, which increases the conductivity of dust and ash deposited on insulators, may increase the attenuation in dB by as much as two to one. Later, heavy rainfall may actually reduce line losses by washing the insulators clean. Hoar frost and sleet, or line icing, are known to increase line attenuation radically at PLC frequencies. Attenuation at lower PLC frequencies, in the vicinity of 50 kHz, has been reported to increase between two and four times from these causes, as compared to the fair-weather values of attenuation. At higher frequencies, increases of attenuation from five to eight times have been reported. For this reason, reliable operation of a PLC communication link under conditions of heavy frost or sleet may require a very costly investment in reserve power, especially if higher frequencies must be used.

COUPLING

The necessary isolation of a line section at carrier frequencies used for communication is accomplished by a line inductor connected in series with the line. It is usually tuned to form a line trap, so that it presents a high impedance over a band of frequencies which may be either narrow or wide.

Separation of the power-frequency energy from the carrier-frequency band is accomplished with the coupling capacitor, and the function of the line-tuning unit is to cancel the reactance of the coupling capacitor to achieve optimum impedance matching and power transfer of the carrier signal to and from the transmission line. The line-tuning unit may also tune out the reactance of the coupling capacitor over either a narrow or a wide band of frequencies.

Some proposals have been made for utilizing the impedances of the line trap, the coupling capacitor, the station-bus impedance, and the line-tuning unit all as parts of a filter pair designed to perform the functions of separation and coupling in an optimum fashion.

The two most commonly employed coupling methods, center-phase to ground, and adjacent-phase to phase, are illustrated in Figure 2.12. Pure Mode-1 coupling can be achieved by a more elaborate scheme for coupling to all three phases, but this is costly.

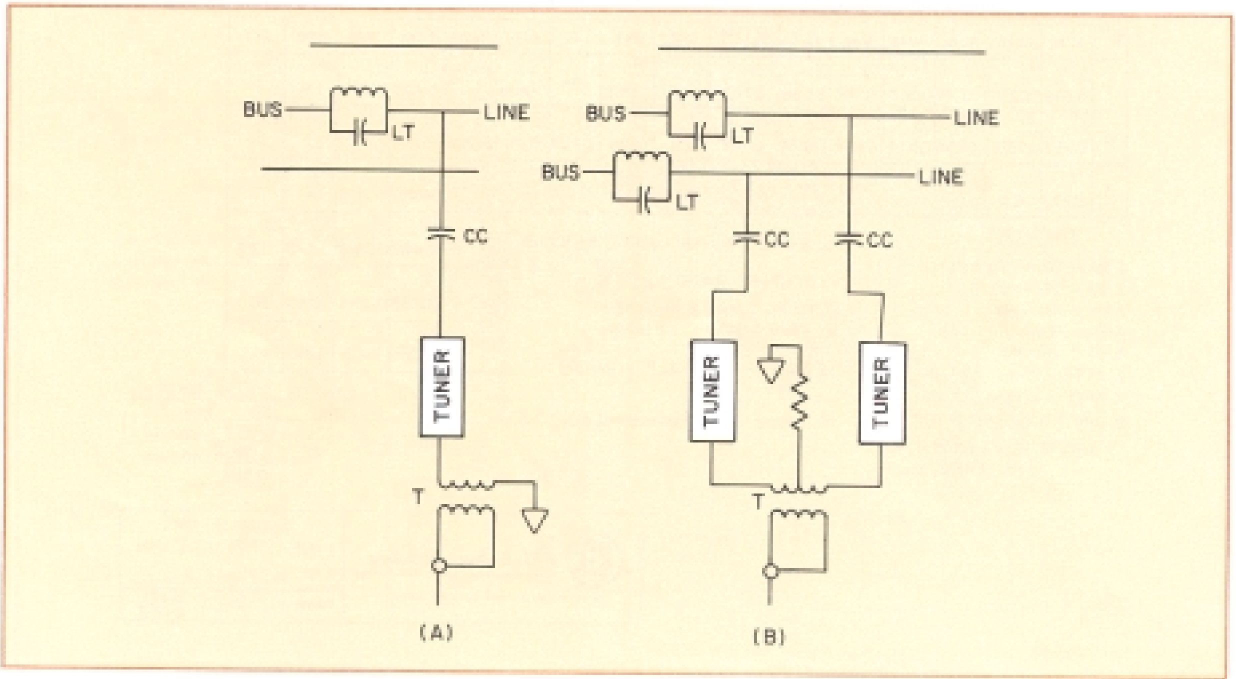


Figure 2.12. Coupling circuits: (a) center-phase to ground, and (b) adjacent-phase to phase.

Coupling Capacitors

The coupling capacitor, CC, Figure 2.12, must withstand the full power voltage of the line, plus additional surges, without breakdown. It serves to block the power frequency from the carrier equipment while presenting low impedance at carrier frequencies. It is available in voltage ratings to 500 kV, or higher, and typically has capacitances from 2 to 12 nanofarads. Sometimes capactive voltage transformers (CVT's) are employed by power utilities for monitoring or control, and these may do double duty as coupling capacitors for carrier communications.

Line Traps

The section of line used for PLC must be electrically isolated, at carrier frequencies, from the station bus or other line sections with a line trap, LT, Figure 2.12. This is a large inductor, capable of continuously handling high values of power-frequency current, and rugged enough to withstand the stresses of higher current surges. It ranges in inductance from about 0.265 mH to about 3 mH. While large inductances are very expensive, smaller values do not normally have sufficient reactance to decouple adequately at lower PLC frequencies; so it is common practice to tune the line inductor with a tuning pack, which may form a resonant trap at one or two specific frequencies for narrow channels, or which may present an impedance of 400 to 600 ohms over a broad band of PLC frequencies. While narrowband traps are largely reactive, broadband traps are largely resistive within the resonant band, and thus they provide assurance against unwanted resonance with unpredictable reactance of the station bus or other line section.

Wideband traps may be supplied either as fixed-tuned or as field-adjustable. The latter may be advantageous where the carrier frequencies to be used are unknown. The adjustable type is useful in this case because a wide range of carrier frequencies may be blocked in groups of smaller, discrete bands. This trap also accommodates present or future changes in carrier frequencies.

Sometimes the station-bus impedance is shunted with an additional bypass capacitor, and an untuned line inductor is employed to form a lowpass filter. This is in parallel with the highpass filter formed from the coupling capacitor and line-tuning unit to form a lowpass-highpass pair coupling to the powerline, as shown in Figure 2.13. This arrangement has low coupling losses over a broad band of frequencies, but it involves more expensive equipment at the coupling site than conventional broadband traps and line-tuning units. Proper choice of components gives a very wide carrier band, low coupling loss, and good isolation from the substation bus. Field tuning is eliminated because there are no adjustable elements, and this enhances reliability where power-system disturbances are severe.

Inductance of the line inductor, 0.67 to 2.0 mH, is high compared to that of a tuned line trap. Its Q is low at carrier frequencies, high at power frequency. Self resonance occurs at a low frequency. These factors give the line inductor an impedance characteristic over a wide frequency band which is superior to that of a wideband trap. Figure

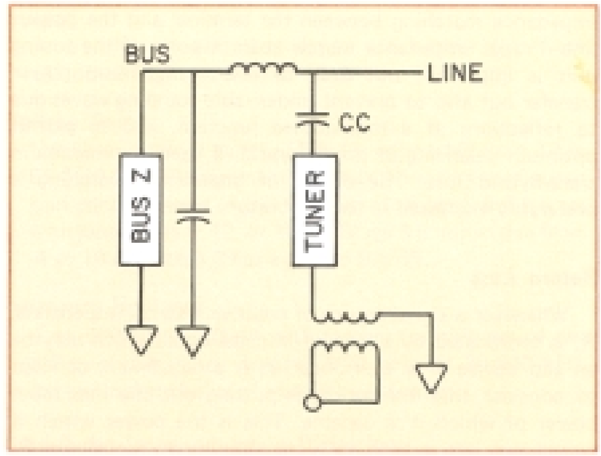


Figure 2.13. Circuit used for lowpass-highpass coupling.

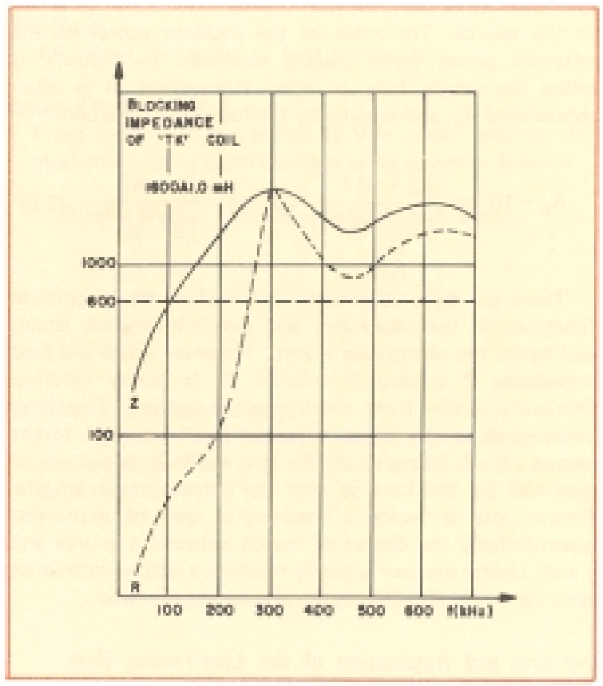


Figure 2.14. Typical impedance and resistance of a 1600-ampere, 1.0-mH coil.

2.14 shows typical impedance and resistance of a 1600-ampere, 1.0-mH coil.

Tuning Units

For efficient coupling at carrier frequencies, it is necessary to cancel the reactance of the coupling capacitor by means of inductance in the line-tuning unit, and to match the impedance of the powerline. Powerline impedances are typically 250 to 400 ohms, phase-to-ground, and 420 to 800 ohms phase-to-phase. Commonly used estimates are 300 ohms and 500 ohms, respectively.

The transformers which are included in line-tuning units are typically provided with several taps to facilitate optimum impedance matching between the terminal and the power-line. Proper impedance match at both ports of the tuning unit is important not only to assure maximum power transfer but also to prevent undesirable standing waves due to reflections at a mismatched junction, and to permit optimum balancing of the rf hybrid, if used, for maximum trans-hybrid loss. The degree of impedance matching is preferably expressed in terms of return loss.

Return Loss

Whenever a signal source of resistive internal impedance, R, is terminated by a load of impedance Z, which may be an end device or a transducer, it is a convenient concept to consider that the source is putting out the maximum power of which it is capable. This is the power which it would put into a load equal to the source impedance, R, and it is referred to as the incident wave. A portion of this incident power will be absorbed by the load or transferred to a load, while the remainder will be reflected, or returned to the source. The ratio of the incident power to the reflected power wave, usually expressed in decibels, is called the return loss, or echo attenuation. It is often abbreviated A_e, and is given by the following Equation:

$$A_e = 10 \log \frac{|Z+R|^2}{|Z-R|^2} = 20 \log \frac{|Z+R|}{|Z-R|} \qquad (2.8)$$

There are three conditions under which the magnitude functions in the numerator and denominator are equal, and hence the return loss is zero. These are when the load impedance Z is zero, is infinite, or is purely reactive. Obviously, under these conditions the load can absorb no power and so the incident power is all returned to the source. If Z is equal to R, the incident power is all absorbed by the load so that the return loss is infinite. Return loss is hence a meaningful way of expressing quantitatively the degree of match between a source and a load. Unless the load is purely resistive, a simple impedance ratio cannot express mismatch in a meaningful way.

Features and Application of the Line-Tuning Unit

The line-tuning unit, like the trap, may be resonant with the coupling capacitor at either one or two specific frequencies, or across a broad band of carrier frequencies. Frequently, the coupling capacitor and line-tuning unit form a highpass filter. The filtering required is very simple, but the tuning unit will have to pass carrier signals at powers up to 100 watts, and it must usually be equipped with a drain coil, surge protector, safety shorting switch, etc. The tuning unit is normally located close to the coupling capacitor in the switchyard, and it must be weatherproof. For phase-to-phase coupling, it is common to employ two line-tuning units, driven by a suitable push-pull transformer.

S/N RATIO, SYSTEM LOADING, and OUTPUT POWER

These three subjects are closely interrelated and are, in turn, related to the system noise levels and losses discussed earlier. Once the noise levels and system losses, including coupling and shunt losses, are known or estimated, the S/N ratio at the receiving end can be used to determine the necessary output power from the transmitter. This is the power necessary just to compensate for all losses in the system plus that needed to raise the received-signal level above the noise level by whatever amount is set by the designer as a desirable S/N ratio.

If the channel serves a single function, such as a voice channel, then the S/N ratio and the output power can be related to the test-tone level representing the average speech power. For speech which is not compandored, the S/N ratio should be 30 to 40 dB for good quality and intelligibility. For example, if the noise in a speech channel is at a level of −20 dBm, then good speech quality can be achieved if the received-signal level is 10 dBm. If line and coupling losses total 25 dB, then the transmitted output power will have to be +35 dBm, or about 3.16 watts, for this voice channel.

If, however, a channel serves multiple functions, such as speech, telemetering, data, and signaling, the situation is more complex. Different types of noise will have different effects upon different types of detectors. Because average noise power is proportional to bandwidth, narrowband tone channels will have relatively less noise power within their bands. Moreover, FSK carriers are inherently less sensitive to noise than are AM-type signals. For such reasons, it is normally possible to operate the tones at levels 10 dB or more below the test-tone level established for speech.

Channel Loading

When computing the loading for a multifunction channel, that is, the numbers and relative powers of the various functional components of the channel, it is necessary to use signal voltages rather than signal powers. Amplifiers and other components will be rated to handle certain peak voltages without clipping and distortion, and the peak voltages of the various signals within the channel may add instantaneously. The ratio of peak to average signal, however, especially for voice, is so great that it becomes both expedient and acceptable to introduce a diversity factor to account for the fact that statistically peaks almost never add simultaneously.

System diversity factor is defined as the ratio of the sum of the individual maximum demands of the various subdivisions of a system to the maximum demand of the whole system. It is normally greater than unity, especially for systems with many subdivisions. For this reason, the practice of rating the system power as corresponding to the sum of the peak instantaneous voltages across the load, which is called the peak envelope power (PEP), is a conservative practice which, for economic reasons, may not always be feasible. Nevertheless, system components will often be rated in terms of PEP, and so channel loading should be computed on that basis.

Equation (2.9) may be used to compute the PEP pf a multifunction channel:

$$P = P_t \; [\; \Sigma \; (N_k \; R_k)\;]^2 \qquad (2.9)$$

where: P is the PEP,

P_t is the test-tone-power level,

N_k is the number of channels of Type k,

and R_k is the ratio of the voltage in each type of channel to the test-tone voltage.

An example will make this clearer.

Assume a system in which the test-tone level is 28 dBm (0.794 watts), and assume the channel is loaded with the following:

(a) One voice channel, for which it is assumed that the peak power will be 3 dB above the test-tone power. This sets R_k at 1.14 for that channel.

(b) Two tone channels with peak power of −10 dB below test tone. Thus, for these channels $R_k = 0.316$.

(c) Six tone channels with peak power of −6 dB below that of the test tone, so that $R_k = 0.5$.

Then, for this loading, the PEP for the channel will be, using Equation (2.9):

$$P = 0.794\; [1.41 + (2 \times 0.316) + (6 \times 0.5)]^2 = 20.23W \qquad (2.10)$$

This may be compared to the earlier example in this section, in which test-tone level required was 3.16 watts which, when used in (2.9), would give a PEP of 80.54 watts, even though the rms power is only 8.53 watts. This indicates the extreme economic penalty involved in designing conservatively for PEP.

Although voice channels may have the most stringent requirement for a good S/N ratio, it should be noted that the apparent S/N ratio of such a channel may be improved by as much as two to one, in dB, by using a compandor. That is, the required S/N ratio may be reduced by a factor of two and, thereby, substantially reduce the power level needed at the transmitter. See Compandors, page 8.

Tones used for protective relaying have demanding requirements with regard to dependability and security. They must function accurately while faults are on the line, at which time both line losses and noise levels are high. For this reason, systems often will elevate or "exalt" line-protection signals when transmitting a trip signal, while simultaneously suppressing all nonessential communication functions. This practice will be necessary to conserve costly PLC power where line loss and/or noise are high.

PLC terminals used for two-way communication normally employ rf hybrid circuits for two-wire to four-wire conversion. Since the usual 3-dB hybrid loss would involve a loss of one half of the transmitter's power, the hybrid is customarily skewed to reduce the loss in the sending direction to about 0.5 dB. This results in about 12 dB of hybrid loss in the receiving direction; but since both signal and noise are attenuated equally there is no deterioration of the S/N ratio.

The two-wire impedance of the hybrid should be specified to match the characteristic impedance of the cable used to connect the terminal to the line-tuning unit. Proper match is necessary if the hybrid is to be balanced for maximum trans-hybrid loss, which will commonly be 35 to 45 dB.

INSULATED SHIELD WIRES

The overhead shield wire(s) on a powerline can be used as a transmission medium by insulating the conductors from the towers. These are usually isolated with an insulator having an air-gap breakdown of 15 to 25 kV. Thus, the original purpose for the shield wires as lightning protection is not compromised. A benefit from using overhead shield wires for communication is the reduction of power-frequency drainage currents induced in them. The resulting power saving may be significant for a shield-wire pair in which the conductors are properly transposed.

Insulating the shield wires causes a slight increase in the zero-sequence impedance of a powerline circuit, and this can cause a corresponding increase in overvoltages associated with line-to-ground faults. In most systems this can be neglected, although in systems with high ground impedance it may be significant.

Terminating Equipment

A suitable coupling device must be used at each shield-wire terminal to provide a drainage path to ground for the induced power-frequency influence, and for lightning currents, while coupling the carrier signal to the shield wires.

The cost of the coupling equipment required for an insulated-shield-wire system is less than for phase-wire coupling because the high-voltage insulation of coupling capacitors and wave traps is omitted. Thus, the cost of shield-wire coupling equipment remains relatively low and constant as a function of the voltage on the system. The shield-wire channel also has an advantage in not requiring a powerline outage when maintenance is required, whereas this is always needed for major maintenance of phase-wire equipment.

Conductors and Insulation

Conductors used on a new insulated-shield-wire channel are usually aluminum-jacketed steel cables. Similar insulation is sometimes added to existing steel wires. The characteristic impedance of a single overhead insulated-shield-wire line is approximately 500 ohms. For a balanced pair, it is about 900 ohms.

Transpositions

Whenever two insulated shield wires are used as a pair, transposition is necessary for saving 50- or 60-Hz power, to provide balance in the communication circuit, and to make the termination requirements less difficult. The objective in selecting locations for transposition is to equalize the induced power-frequency current in the two wires. This is done by equally dividing the distance over which each wire occupies each position in the line's configuration. The maximum distance between transpositions will be determined by how much voltage should be permitted to exist on the shield wires during heavy powerline loading. Depending on line current, and on user practice, guidelines limiting this distance to values from 6 to 30 miles have been used.

Performance

Flashover across shield-wire insulators will cause the carrier-frequency attenuation of the circuit to increase, but usually not enough to cause a communication outage. The duration and the amount of the increase depend on the cause of the flashover and the momentary power-circuit conditions which accompany it. For example, where lightning causes a flashover which occurs only on the shield wires, the added loss will be small and its duration typically of the order of one millisecond, or less. Measurements have indicated an increase in loss ranging from 1 to 8 dB, depending on carrier frequency, during flashovers near the end of a line. Flashovers which accompany a powerline fault will last until the fault is cleared, typically 60 milliseconds, or longer.

During wet weather there is approximately 20-percent increase in attenuation. Insulated shield wires are more susceptible to increased losses due to frost than are phase wires, for they do not have significant current-induced heating. Moreover, ice bridges are sometimes formed over the relatively small insulators. Complete communication-circuit outages have been experienced during severe icing conditions.

Terminal equipment used on shield-wire channels can be identical to that used on phase-wire channels. Most of the transmitters used currently have one watt or greater output.

A frequency range from 8 to 500 kHz is technically feasible for shield-wire channels. Because high noise and interference are present at low frequencies, the coupling equipment must be designed to provide a bandpass or highpass filter for the carrier signal. A cutoff frequency of about 5 kHz, and attenuation of at least 40 dB below this frequency should be used.

Insulated-shield-wire channels are used for voice, supervisory control, alarm, and telemetry functions. Some protective-relaying channels have been applied, but there is limited experience with this application.

Bibliography

The following references on the subject of insulated-shield-wire channels may be of interest:

"Application and Operating Experience of Carrier Communications Over Insulated Static Wire", T. D. Wood, Jr., Paper No. 31-CP-66-41, IEEE Winter Power Meeting, New York, N. Y., January 30-February 4, 1966.

"Use of Insulated Ground Wires on a Transmission Line for Communication Channels". G. E. Farmer, IEEE Transac. PAS, No. 69, pp. 884-891, (Dec. 1963).

"Protection of Communication Circuits Serving Electric Power Systems", C. L. Roach and G. Y. R. Allen, CIGRE Report No. 308, 1966.

Guide for Power Line Carrier Applications, Proposed IEEE Standards Project P643/D1.

INTRABUNDLE CHANNELS

Bundled phase conductors can provide a means for sending PLC signals on a single phase of a powerline without the need for a ground-return path. The approach involves the transmission of signals on two or more conductors within a multiconductor phase bundle in which the individual conductors are insulated from each other. At this time, applications have been limited to experimental lines.

Theoretical and experimental work has been carried out on such channels, and an experimental intrabundle line has been operated in the USSR (1).

Application

Intrabundle channels are essentially untried, and they involve the use of several unproved techniques. Experience is lacking with line-coupling methods, with performance of insulated separators, and on behavior of faults. Loss caused by icing is an important problem.

The technique, however, has several significant advantages, including:

(a) Lower fair-weather line losses than conventional carrier channels.

(b) Large crosstalk attenuation between intrabundle channels.

(c) Higher density application of channels.

(d) Less interference with or from radio services in the PLC band.

It is believed that the intrabundle channel may become an important technique in future application of PLC, particularly in ice-free areas.

(1) "Measurement and Calculation of Intrabundle HF Communication Paths", E. E. Brestkina, et al, (USSR) Paper 35-03, CIGRE International Conference on Large High Voltage Electric Systems, Paris, France, Aug. 21-29, 1974.

Chapter 3

TYPICAL SYSTEMS

INTRODUCTION

The PLC terminal provides the transmitting and receiving channels required for all kinds of communication functions using this medium of transmission. These include voice, signaling, telecontrol, telemetering, and protective-relaying systems, all of which operate in the voice-frequency band. The terminal equipment provides these channels by using frequency translating equipment, power amplifiers as needed, and means for coupling to the powerline.

MODULATION

Choice of modulation plans for a PLC system depends upon both the detailed nature of the transmission line to be used and on the kind of communication service needed. In low-frequency (4-60 kHz) systems using openwire lines or insulated shield wires as the transmission medium, single-stage modulation, using just one frequency translation, is often most economical.

In PLC systems where good selectivity is needed to utilize the available frequency spectrum most effectively, double-stage modulation, using two frequency translations, is used most often.

It will be noted that there is no predisposed modulation plan which is basic to RFL's design for Series 65 SSB equipment. The designer is free, therefore, to choose as many or as few channels as needed, and these may be disposed at any frequency in either or both the intermediate-frequency band and the high-frequency band. This freedom of choice will tend to simplify the planning and design of any PLC communication system. It also offers an easy choice when planning for future expansion of a system.

Single Translation

A single-channel, low-frequency modulation plan is shown at Figure 3.1 (A), in which the lower sideband is transmitted with a suppressed carrier at 20 kHz. The bidirectional arrows at points of frequency change indicate that the system is receiving when a Model 65 DEMOD is used, or is sending when a Model 65 MOD is used as the translator. Figure 3.1 (B) shows an identical system arranged for use of the upper sideband. In either case, the unused sideband, resulting from the modulation process, is suppressed with a filter.

Figure 3.1 (C) shows the same system, to which has been added a second channel to create a two-channel system. In this arrangement, the audio information from Channel A is carried on the lower sideband. Channel-B audio is carried on the upper sideband of the same carrier.

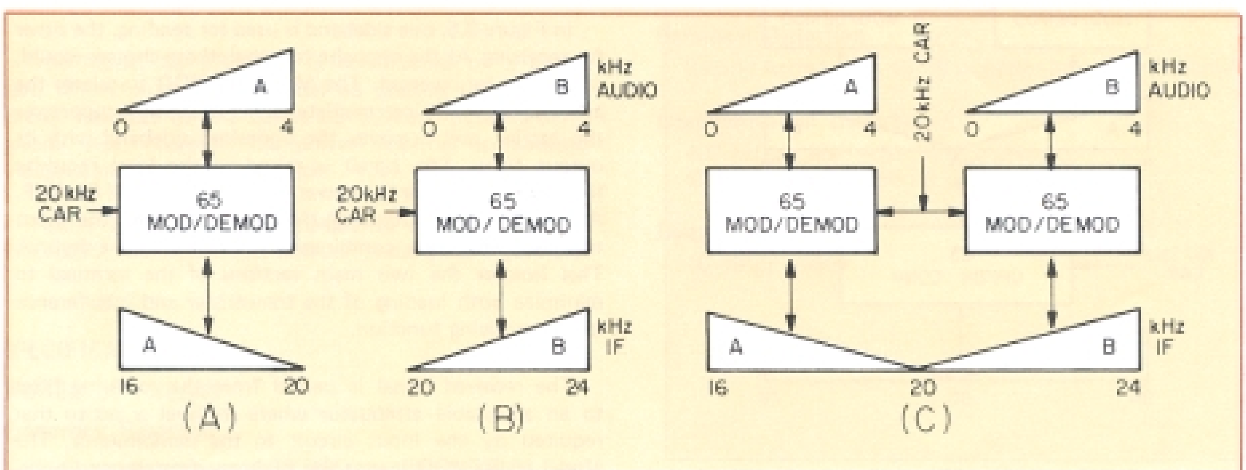


Figure 3.1. Single-frequency translations: (A) lower sideband, (B) upper sideband, and (C) two-channel system.

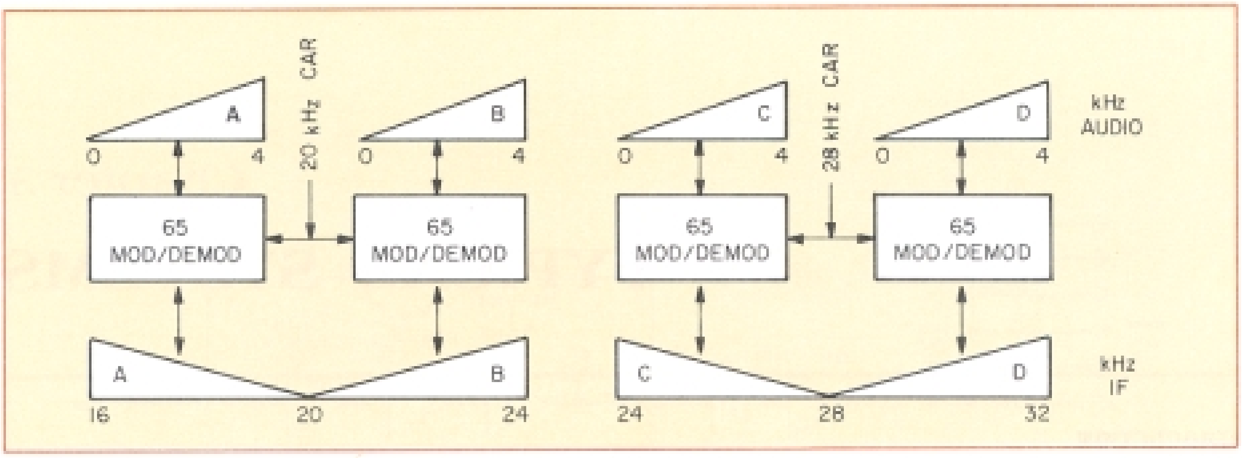


Figure 3.2. A four-channel system using single-frequency translation.

Figure 3.2 shows a four-channel system using a single frequency translation. This is simply two of the systems shown in Figure 3.1 (C), each of which has a different carrier frequency. A single pilot carrier in either of the channels is sufficient for controlling both carriers for demodulation at the receiving terminal.

Double Translation

Figure 3.3 outlines a two-channel system using double translation. It will be a sending or receiving terminal depending upon choice of modulator and an upconverter for sending or of demodulator and downconverter for receiving. If only a single-channel system is wanted, audio-channel A and one modulator or demodulator are omitted, and filters, if needed, are used to suppress unwanted sidebands. Typical frequencies are shown, but it is apparent that any of these systems may be used with frequencies at any place in the frequency bands of the modulator, the demodulator, and the up and down converters.

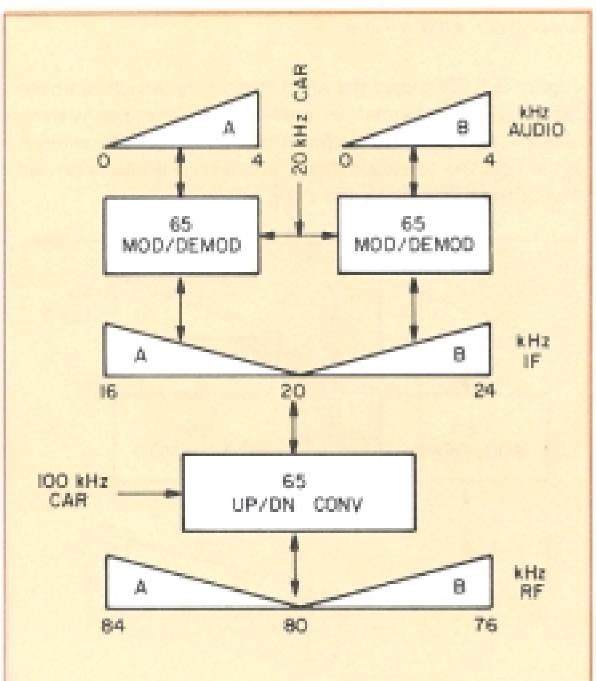


Figure 3.3. A two-channel system using double-frequency translation.

The economical utilization of terminal equipment becomes apparent in Figure 3.4, which is simply two of the systems shown in Figure 3.3, making a four-channel system in which only one converter is needed for the high-frequency translation. This concept can be extended to provide several more channels, all passed through just one converter.

LOW-FREQUENCY SINGLE-CHANNEL SYSTEM

A single-channel, low-frequency, two-way (duplex) PLC terminal is outlined on Figure 3.5. Identical equipment is required at the opposite terminal. The Models 65 MOD and 65 DEMOD provide for any carrier frequency between 4 and 60 kHz. If there are no constraints, the system usually is operated at a suppressed-carrier frequency between 20 and 30 kHz.

In Figure 3.5, one sideband is used for sending, the other for receiving. At the opposite terminal, these choices would, of course, be reversed. The Model 65 MOD translates the audio signal to the intermediate carrier frequency, suppresses the carrier, and removes the undesired sideband with its output filter. The signal is raised to the level requisite for transmission over the powerline with the Model 65 AMP. At the line, the output of the transmitter and the input to the receiver are combined with filters and a hybrid. This isolates the two main sections of the terminal to minimize both loading of the transmitter and interference to the receiving function.

The received signal is passed from the receiving filter to an adjustable attenuator where its level is set to that required by the input circuit to the demodulator. The Model 65 DEMOD inserts the suppressed carrier and translates the signal to the audio band.

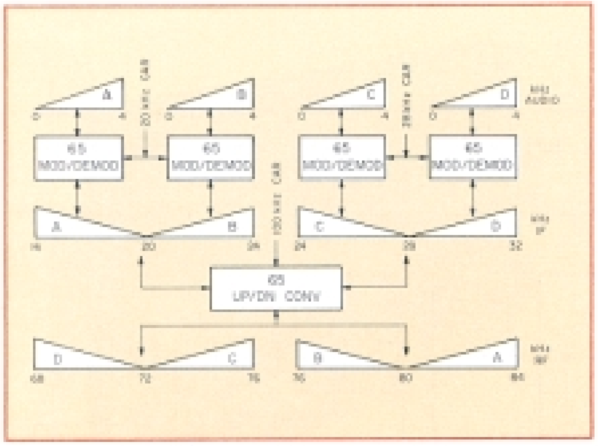


Figure 3.4. A four-channel system using double-frequency translation.

The Model 65 COMP Compandor, shown on Figure 3.5, contains both a compressor for transmitting and an expander for receiving speech signals. Because powerline noise is greater at lower frequencies, the Model 65 COMP is recommended for this system if used for telephony.

The Model 65 FS SIG provides for sending and receiving the announcement of a call. The signal directed to the opposite terminal is entered on the M lead. A call signal from the other terminal is received on the E lead. The signaling tone used is frequency-shift keyed, and, therefore, is always present.

The receiving side of each terminal is equipped with automatic-gain control (AGC) having a dynamic range of more than 40 dB. An AGC circuit is included in the af section of the demodulator, and control is provided from the optional AGC circuit of the signaling transceiver. The out-of-band signaling tone is used as a pilot for the AGC. This compensates for changes in attenuation of the powerline, such as those caused by adverse weather.

The af hybrid provides for transformation of the four-wire af ports of the transmitter and receiver into a two-wire drop suitable for connection with a standard telephone line, PBX, or local instrument.

If required by the system, a Model 65 SYNC Synchronizer may be added to synchronize each receiving unit with the carrier frequency of its companion sending unit at the opposite terminal, in which case the carrier will be transmitted at reduced level, rather than completely suppressed.

HIGH-FREQUENCY SINGLE-CHANNEL SYSTEM

A single-channel, high-frequency PLC terminal is outlined in block-diagram form on Figure 3.6. The terminal is essentially the previously described low-frequency terminal, to which has been added the necessary frequency translators and auxiliary equipment. For this single-channel system, upper and lower sidebands of a suppressed carrier may be used for sending and receiving.

On the transmitting side, the SSB output of the 65 MOD is fed to the Model 65 UPCONV, where it is translated to an RF carrier frequency. A bandpass filter is used to remove the unwanted sideband. One or more amplifiers, as necessary, are used to raise the output level to overcome the losses and attenuation of the powerline. An additional filter at the output rejects spurious signals from the amplifier.

The rf hybrid and filter are used to isolate the high-power signal of the transmitter from the sensitive receiver input. The hybrid is usually skewed to minimize power loss in the transmitting direction at the expense of greater loss in the receiving direction.

The signal received is attenuated as needed to fit the input sensitivity of the Model 65 DNCONV. This module translates the frequency of the input signal to the operating range of the Model 65 DEMOD. From this point, the receiving process is identical to that described for the low-frequency system.

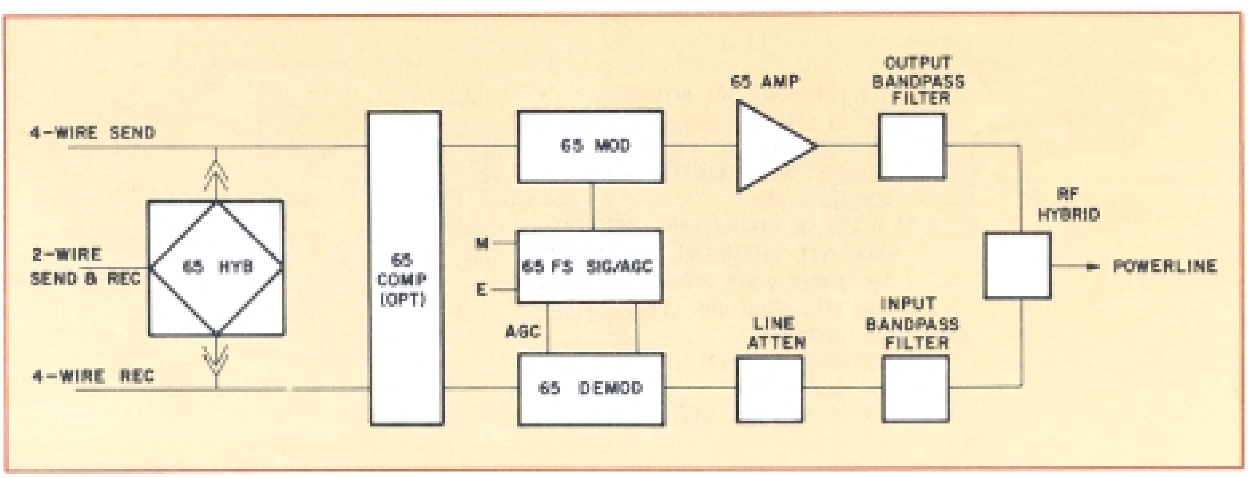
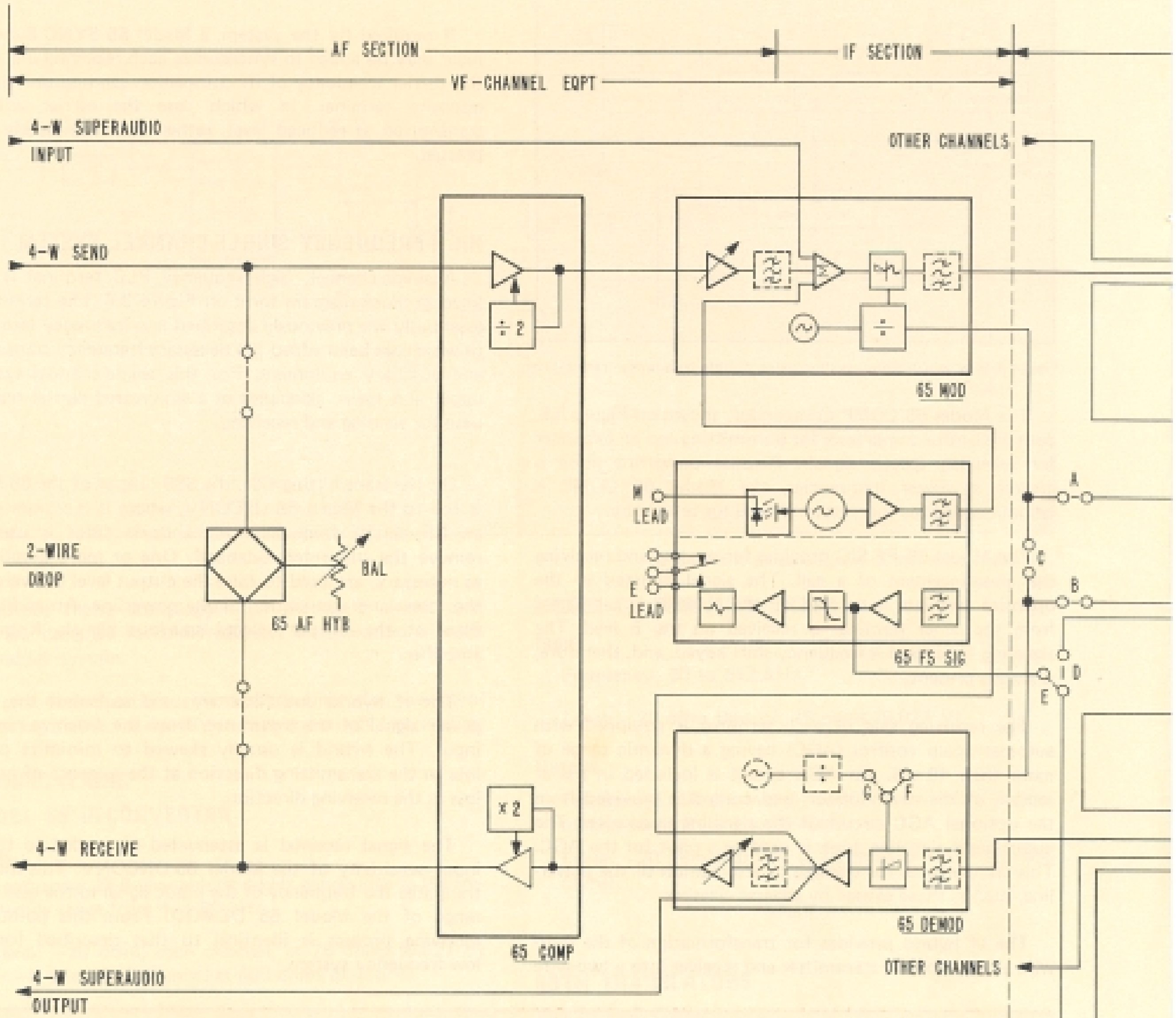


Figure 3.5. A single-channel, low-frequency, duplex terminal.



NOTES

1. DOTTED COMPONENTS ARE OPTIONAL.
2. TO USE 65 SYNC, USE JUMPERS A, B, AND F.
3. TO OMIT 65 SYNC AND RECEIVE ON SAME FREQUENCY AS TRANSMITTER, USE JUMPERS C AND F.
4. TO RECEIVE ON CARRIER FREQ. DIFFERENT FROM THAT TRANSMITTED AND WITHOUT 65 SYNC, ADD JUMPER G AND OPTIONAL OSCILLATOR.
5. TO USE SYNC. PILOT FOR LEVEL CONTROL, USE JUMPER D.
6. TO USE SIGNALING PILOT FOR LEVEL CONTROL, USE JUMPER E.
7. TO USE 65 AGC, OMIT JUMPER H.

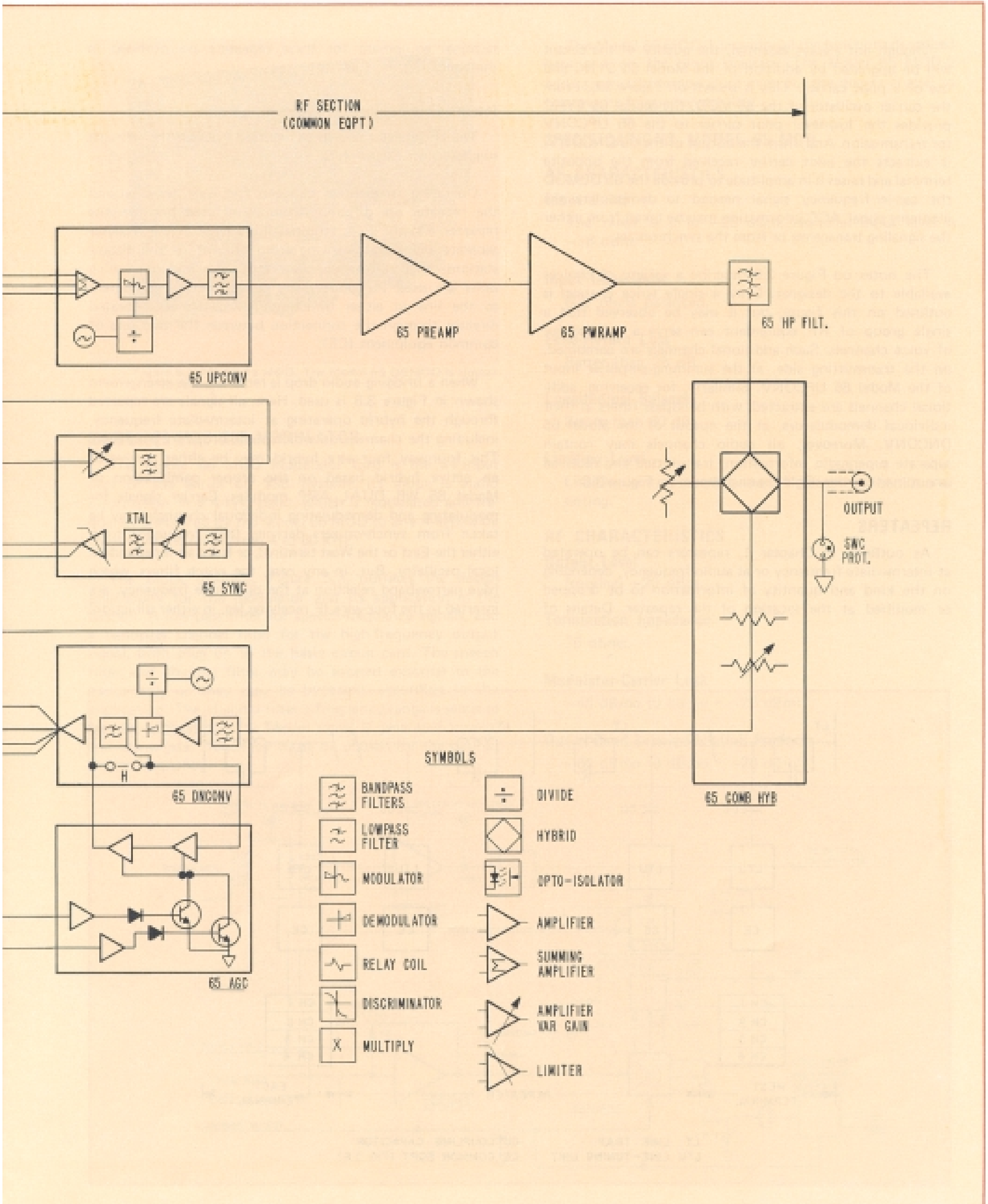


Figure 3.6. A single-channel, high-frequency duplex terminal. Provisions for additional channels are indicated.

Though not always essential, the quality of the circuit will be upgraded by addition of the Model 65 SYNC and use of a pilot carrier. This is shown on Figure 3.6. From the carrier oscillator of the 65 MOD, the Model 65 SYNC provides the low-level pilot carrier to the 65 UPCONV for transmission. And, from the output of the 65 DNCONV, it extracts the pilot carrier received from the opposite terminal and raises it in amplitude to provide the 65 DEMOD the carrier-frequency signal needed to demodulate the incoming signal. AGC information may be taken from either the signaling transceiver or from the synchronizer.

The notes on Figure 3.6 describe a variety of choices available to the designer. Only a single voice channel is outlined on this figure, but it may be observed that a single group of RF equipment can serve a multiplicity of voice channels. Such additional channels are combined, on the transmitting side, at the summing-amplifier input of the Model 65 UPCONV. Similarly, for receiving, additional channels are extracted, with bandpass filters at their individual demodulators, at the output of the Model 65 DNCONV. Moreover, all audio channels may contain separate superaudio information, transmitted and received as outlined for the single channel shown on Figure 3.6.

REPEATERS

As outlined in Chapter 1, repeaters can be operated at intermediate frequency or at audio frequency, depending on the kind and quantity of information to be dropped or modified at the location of the repeater. Details of terminal equipment for these repeaters are outlined in the following.

Intermediate-Frequency Repeater

The IF repeater shown as Figure 3.7 details the terminal suggested on Figure 1.12.

Operating frequencies between the West terminal and the repeater are different from those used between the repeater and the East terminal. Line traps at the repeater separate the signals arriving from or sent to the distant stations. High-voltage coupling capacitors and line-tuning units are used to connect the common (RF) equipment to the line in either direction. All signals are repeated directly through the connection between the two sets of common equipment (CE).

When a bridging audio drop is required, the arrangement shown in Figure 3.8 is used. Here, all signals are repeated through the hybrid operating at intermediate frequency, including the channels on which audio drops are paralleled. This four-way four-wire hybrid may be either passive or an active hybrid based on the proper combination of Model 65 WB DUAL AMP modules. Carrier signals for modulating and demodulating individual channels may be taken from synchronizers deriving their frequency from either the East or the West terminal, or from an independent local oscillator. But, in any case, the notch filters, which have narrowband rejection at the pilot-tone frequency, are inserted in the four-wire IF receiving leg, in either direction,

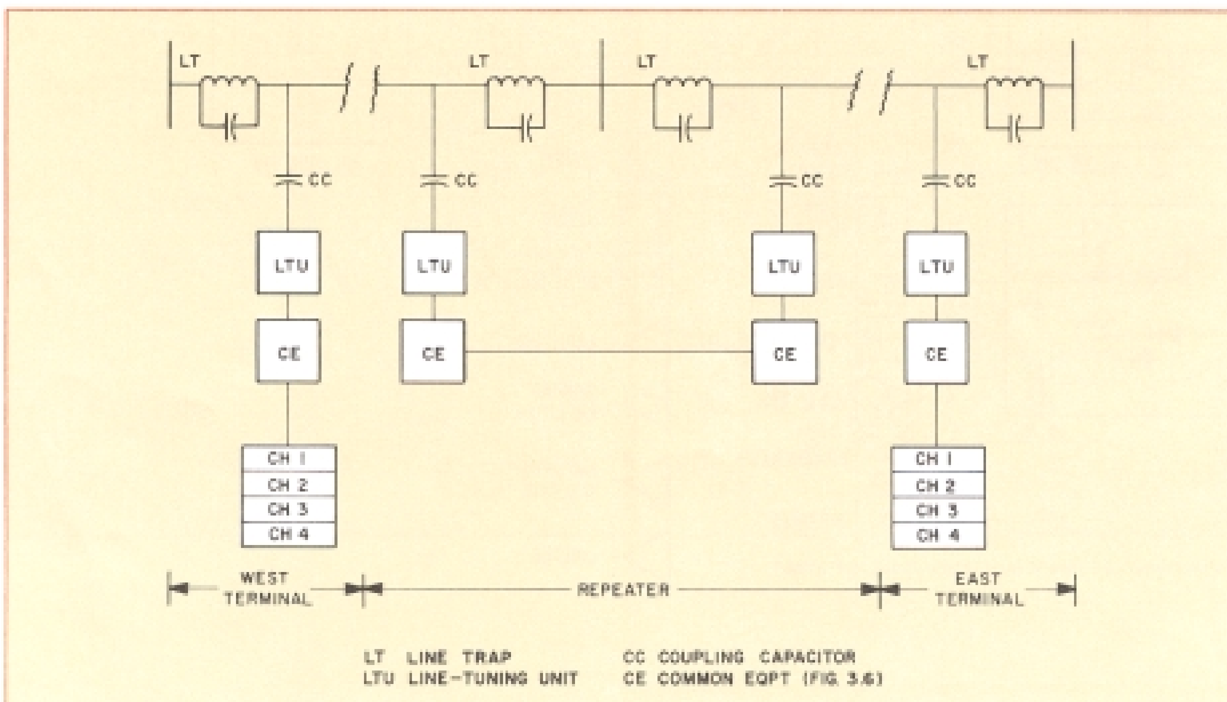


Figure 3.7. Typical IF baseband repeater.

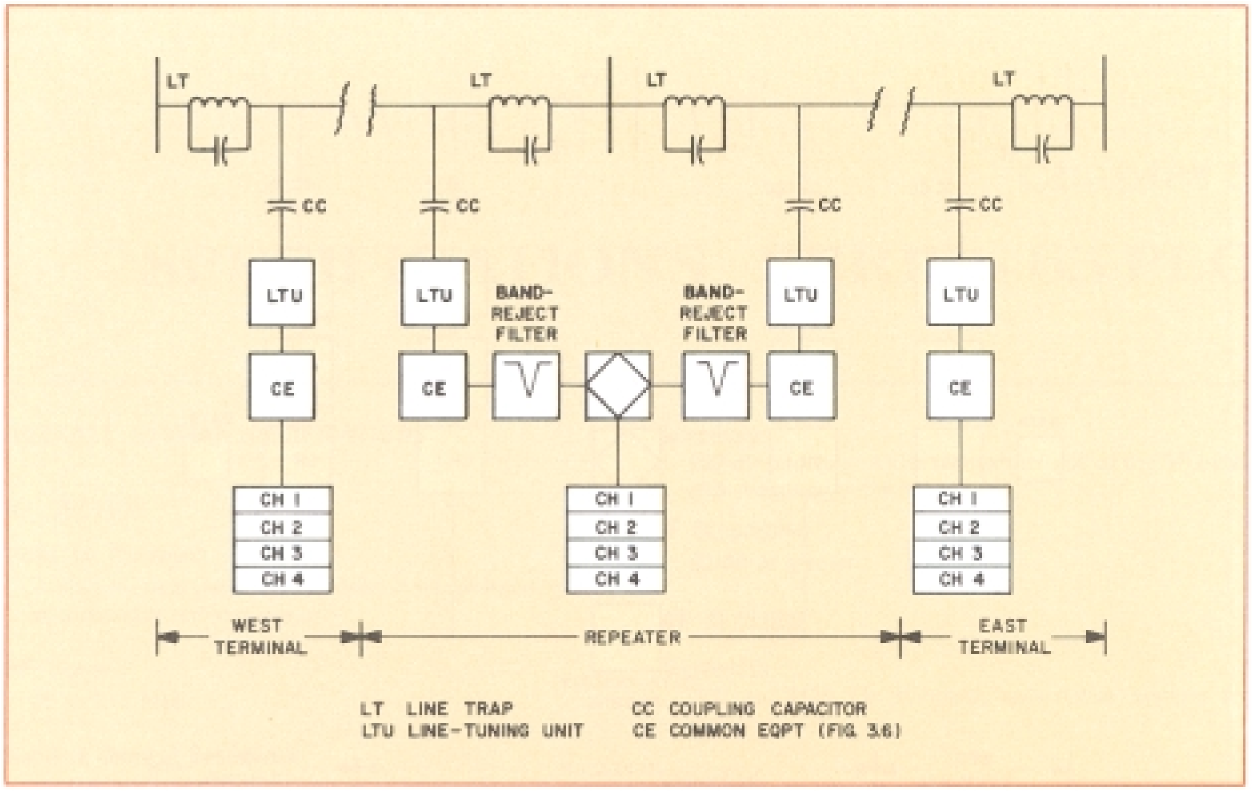


Figure 3.8. Typical IF baseband repeater equipment with audio drop.

to prevent retransmission of the carrier-frequency signal used locally for modulation or demodulation.

Sets of individual channel equipment are required for each channel for which an audio drop is required, and it will be noted that this system does not permit alteration of the content of information on any channel so dropped.

Audio-Frequency Repeater

In Figure 3.9, which details the audio-frequency repeater outlined on Figure 1.14, line-frequency (RF) signals are dropped to audio frequency before being repeated. As for the former repeater, isolating line traps, high-voltage coupling capacitors, line-tuning units, and common RF equipment all are required. In addition, individual channel equipment is required for every channel that is to be repeated. The audio-frequency hybrid indicated for Channel 1 performs the usual function of a hybrid, and one will be required for each channel repeated.

The data-transit filter is a high-pass filter which provides for repeating superaudio tones across the repeating point. If desired, however, the content of both audio and of superaudio information may be altered at a repeating point.

COMPOSITION OF THE SYSTEM

Because the RFL Series 65 is not a rigidly standardized system, offering only one modulation plan and a fixed number of channels, the designer is free to establish the particular PLC system desired, anticipating growth but providing additional equipment only when and if needed. He need select, at one time, no more and no less than the type and quantity of individual circuit modules needed to create or to expand his system.

The various circuit elements, features, and concepts available for use in a system, as discussed in these first three chapters, can be included, omitted, or added later according to need. If requirements change in the future, the system can be adapted easily to supply the new demands.

A typical terminal is shown as Figure 3.10.

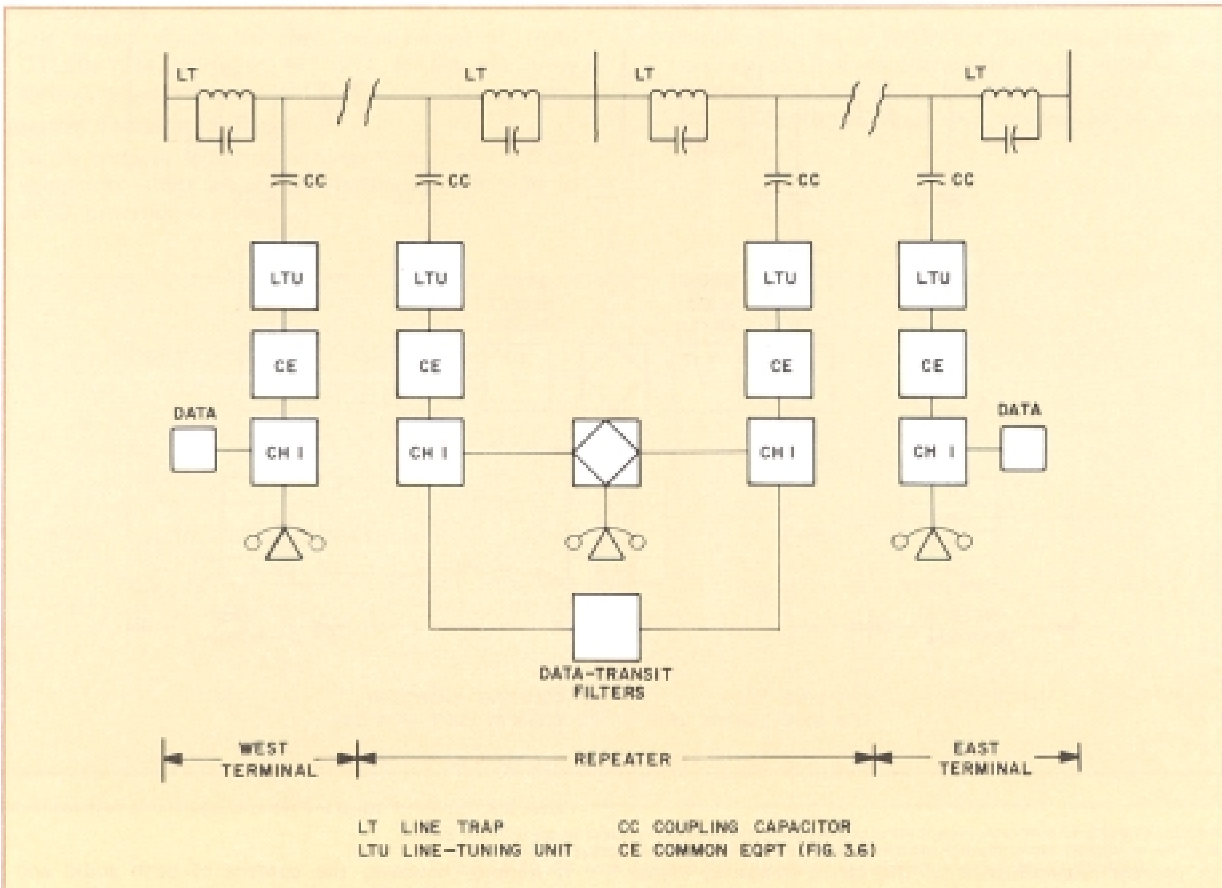


Figure 3.9. Typical AF repeater equipment.

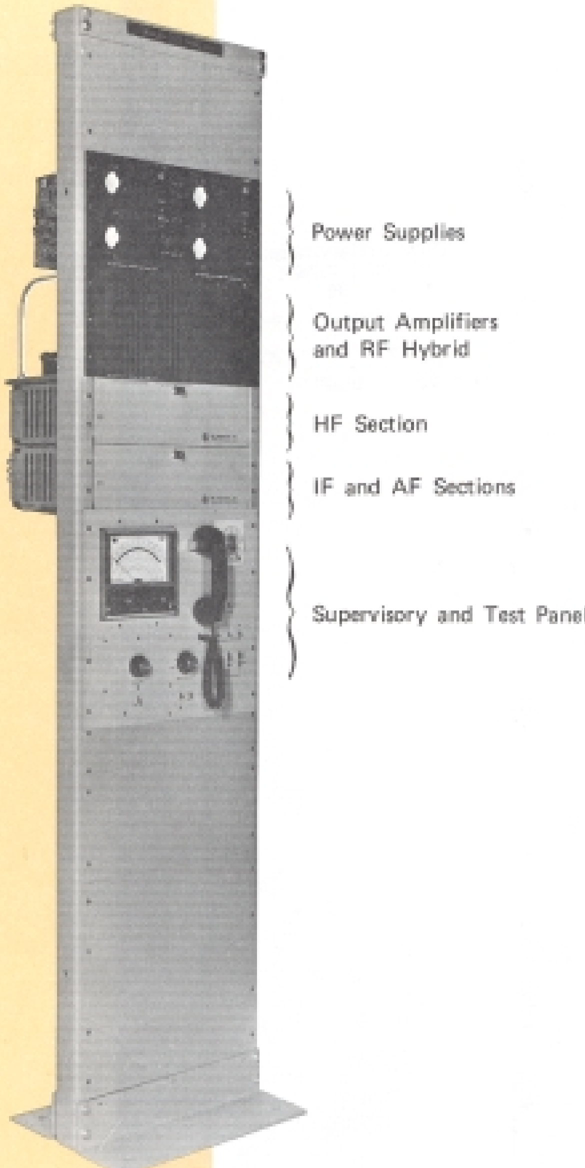
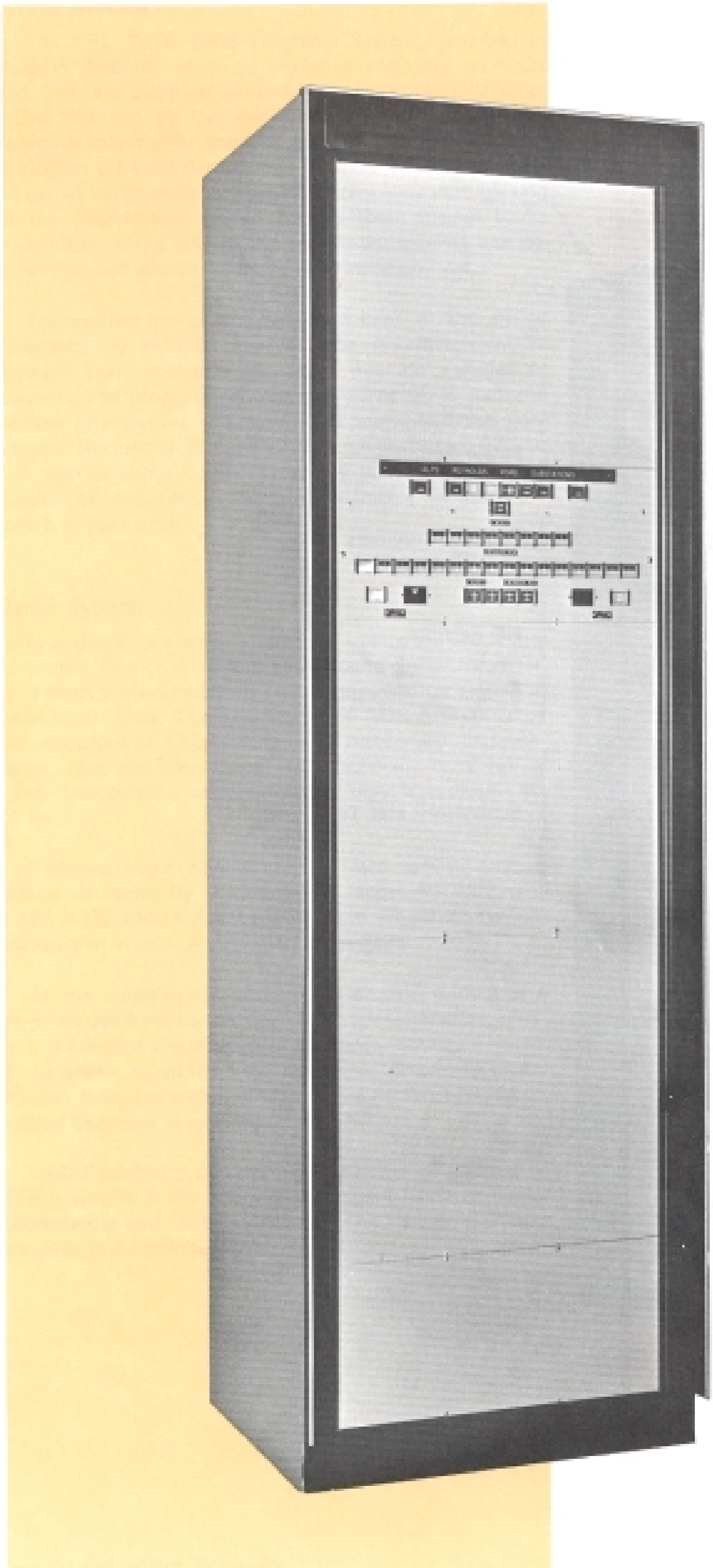


Figure 3.10. Typical PLC terminal.



This RFL Series 66 Time-Division Multiplex System includes 24 points of control and indication and status, with two points of tap change incorporating digital position reporting, all used for supervisory control. It is typical of peripheral equipment that can be connected to a PLC system.

Chapter 4

PERIPHERAL EQUIPMENT

GENERAL

A PLC terminal can be defined as the equipment required to accept one or more two-wire voice channels and operate on them so that a composite signal is delivered to the powerline in whatever form is established by the designer of the system. At the receiving terminal, the reverse operation is effected. In this sense, the terminal presents the user with one or more independent 4-kHz communication channels, each of which may be put to whatever purpose is chosen.

Because RFL specializes in communication systems, it is in a position not only to supply PLC terminals but also to design and supply complete peripheral systems comprising the logical extension of the input and output ports of a PLC terminal. This chapter outlines a few possibilities.

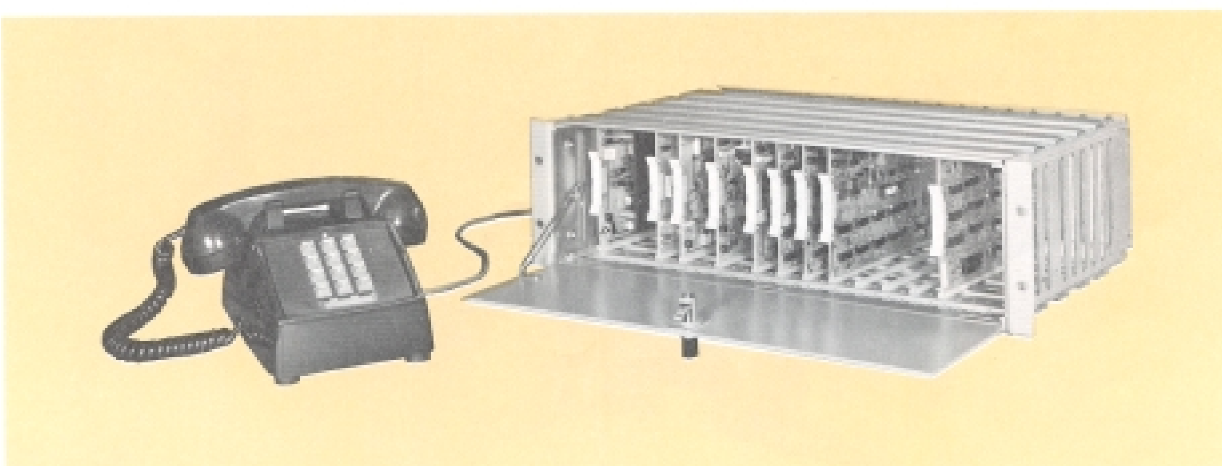
SELECTIVE CALLING TELEPHONE UNIT

The RFL Series 6500 SCTU, Selective Calling Telephone Unit, has been designed to provide two-way selective dialing between telephones at various locations in a party-line telephone communication system.

A standard telephone handset with dual-tone, multi-frequency (DTMF) pushbutton dial can be used at each extension of the RFL Series 6500 SCTU.

A two-digit or three-digit numbering plan may be adopted in the party-line network. If a two-digit numbering plan is used, the party-line network could have as many as 10 terminals; each capable of handling as many as 10 extensions. If a three-digit numbering plan is used, the party-line network could have as many as 100 terminals; each capable of handling as many as 10 extensions.

An optional relay card, available with the terminal can provide remote control and supervision of 10 remote points.



ORDERWIRE SYSTEM

An orderwire is often described as a party-line telephone circuit, in which two or more stations are free to communicate, either individually or as a group, at will. This system of communication has been found to be especially convenient in supervisory-control work, such as management of a powerline, where two or more operators must function cooperatively to achieve a desired goal. RFL's PLC terminals often are equipped with orderwire facilities.

Although described generally as a party-line system, orderwire circuits frequently are set up with dual-tone, multi-frequency (DTMF) signaling in which up to 999 stations can be called selectively. An optional privacy feature permits only the called station, or stations, to be connected to the system. To prevent the originating station from tying up the system for too long a period, an optional timeout circuit will release the system to other callers after a predetermined time. A voice-activity detector may be added to the system to prevent a station inadvertently left off-hook from tying up the system. If desired, dial tone, busy tone, call tone, and timeout tones can be provided. Sidetone, which also provides audible recognition of transmitted DTMF tones, is standard.

E and M signaling, described earlier, is also available.

All RFL orderwire systems are suitable for use in "speech-plus" systems, outlined in Chapter 1.

Figure 4.1 shows representative orderwire terminals. The station at the top includes a keypad for DTMF calling to other stations in the system. Calls received at this station are announced through the loudspeaker.

At center of this illustration is a simple station, suitable for a subordinate remote point at which calls are originated with a single pushbutton.

At the lower part of Figure 4.1 is an orderwire control center at which not only is voice communication used but also a number of different sources of signals can be selected for connection to a variety of different circuits.

Details of orderwire systems are given in Bulletin 6850-LOW, which may be requested.

PROTECTIVE RELAYING

Control signals for protective relaying are almost universally carried by PLC systems for protection of the powerline against short circuits and other overloads. The RFL Series 6745 equipment comprises a system specifically designed for carrying such signals in a format that permits optimizing the necessary tradeoff between dependability and security always required. It is a completely self-contained, solid-state, dual-subchannel system.

Physical appearance of the equipment is similar to that of the PLC equipment, for the same chassis and style of plug-in circuit board are used.

Frequency assignments for the system are taken from a group chosen to operate in the voice-frequency band of the PLC system. One or more protective-relaying circuits may occupy a complete PLC channel, or they may be multiplexed with other signals.

For details on this and other protective-relaying equipment, request additional information from RFL.

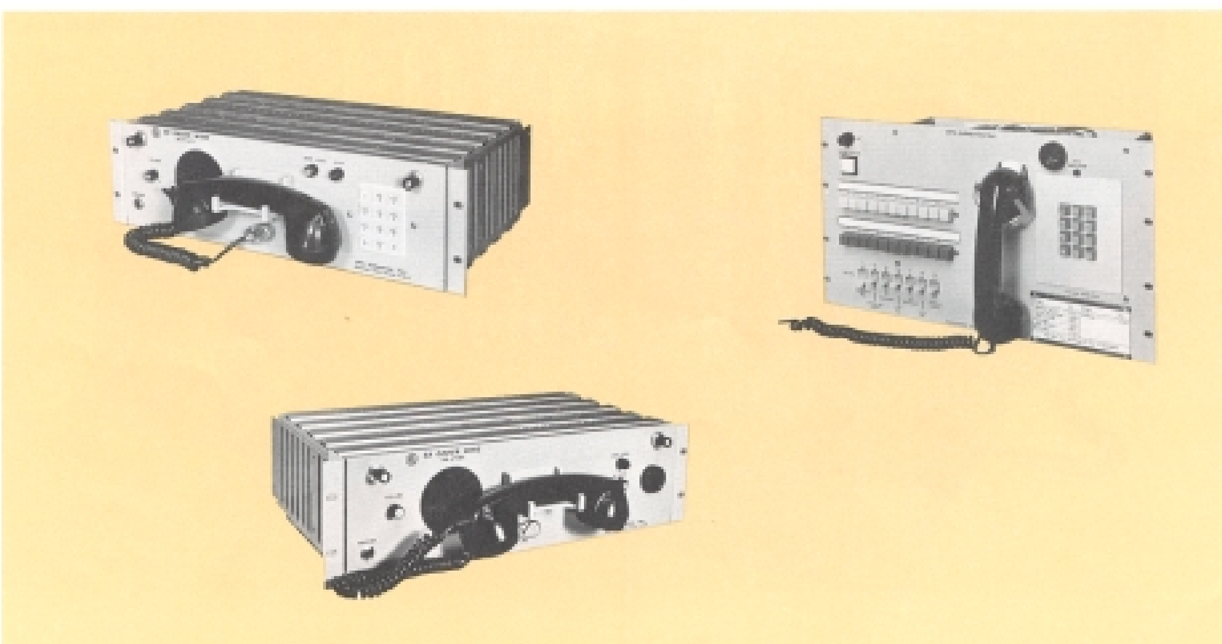


Figure 4.1. Typical orderwire terminals.

VOICE–FREQUENCY CARRIER EQUIPMENT

Series 6850 equipment comprises a group of modules used to provide for multiplexing a quantity of data circuits on a single PLC voice channel by using audio-frequency carrier, often called "tones". Depending upon speed of transmission, up to 27 such channels can be multiplexed on a single voice channel. Bulletin SB6850 is available for details.

Transmitters and receivers have been designed for exceptionally low distortion. They can be used for either two-frequency or three-frequency FSK operation, for amplitude keying (AM), or for both simultaneously. They operate up to 1200 baud in the voice band, and can be operated to 4800 baud at higher frequencies.

Supplementing the basic transmitters and receivers, the series includes modules such as a carrier detector responding not only to absolute level of signal but also to S/N ratio. A line-interface module contains appropriate amplifiers for fanning several inputs or outputs to or from a single line, so that it functions as an active bridge. A bandpass-amplifier module makes it possible to insert or extract a single multiplexed signal in a group. And a line-termination unit can provide specific interfaces needed between signal-processing units and the communication link.

This equipment provides the basic transmitting and receiving terminals for such information as analog or digital telemetry, computer linkage, time-sharing, data collection and transmission, telegraph channels, and supervisory-control signals. It will supply these functions either independently or as a component of a speech-plus system.

VOICE-FREQUENCY CARRIER TELEGRAPH

A unique element of the Series 6850 System, the Model 68A TX/RX is a transceiver (combined transmitter and receiver) designed especially for data applications. Bulletin SBVFCT is available with details. A PLC channel can be the communication medium.

In a single, compact module, this transceiver provides for a wide variety of data-transmission requirements including telegraphy, data, speech-plus-data, supervisory control, and alarms. Its flexibility in digital interfaces provides for either neutral or polar telegraph loops, polar, EIA/CCITT positive or negative logic, TTL, DTL, or CMOS digital interfaces.

Flexibility of assignment, or reassignment, of operating frequencies is achieved by using separate plug-on filter cards for transmitter and receiver. Intermixing channel speeds and bandwidths is possible due to the selectivity and nonloading characteristics of these filters.

Modular printed-circuit boards for plug-on assemblies provide for customized selection of optional features. These are available not only for system requirements, such as polar or neutral telegraph loops, filter selectivity, low-level digital interfaces, request-to-send and clear-to-send delay, carrier detector and amplification, but also for diagnostic and monitoring needs such as v-f carrier and dc jacks, carrier-detector indicator, and received-data indicator.

Figure 4.2 shows two views of the Model 68A TX/RX.

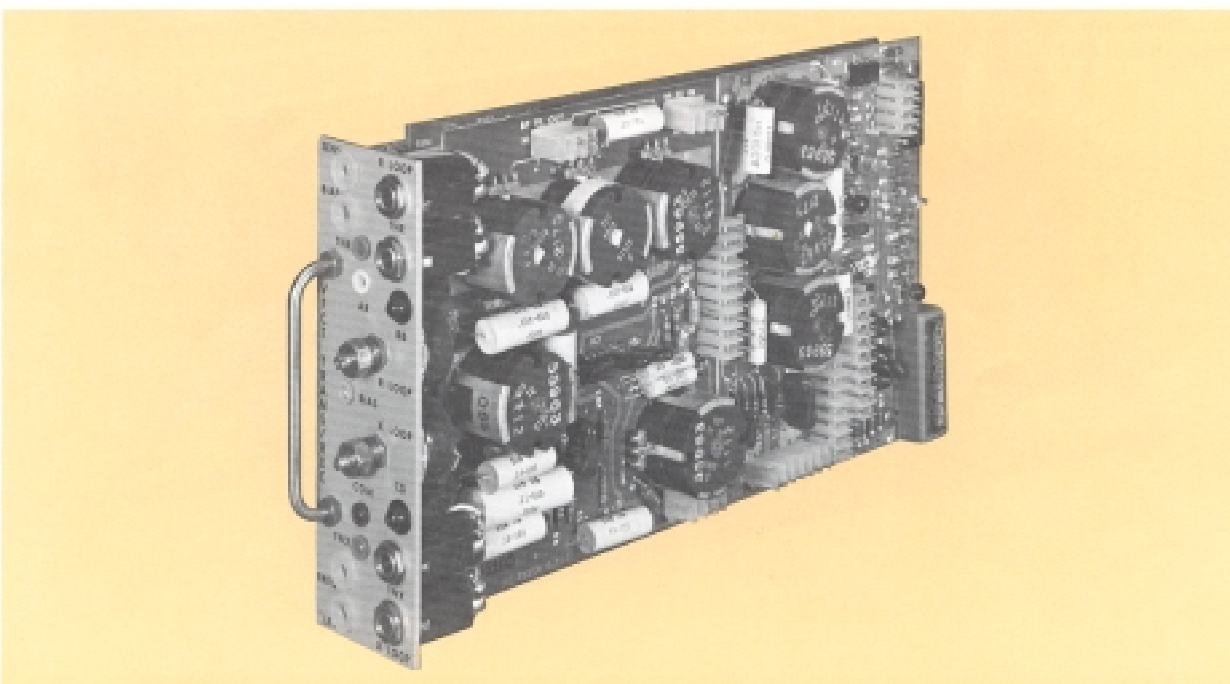


Figure 4.2. Model 68A TX/RX with plug-on cards.

SUPERVISORY-CONTROL SYSTEMS

Supervisory-control systems are used to control and monitor the status of such items as valves, pumps, and circuit breakers. Digital telemetering may be added for transmission of such parameters as voltage, current, power, phase, pressure, flow, level, and temperature. The method of transmission most frequently used is time-division multiplexing of data on a voice-frequency carrier which may be carried on a PLC system.

Using the same physical arrangement as the RFL Series 65 PLC equipment, the Series 66 Time-Division Multiplexing System (TDMS) has been designed specifically for use in supervisory-control and data-acquisition systems. The system permits tradeoffs to be made between security and speed of operation; and this choice, combined with the use of an efficient code, enables a designer to optimize the system to his requirements. The multiplexed signals at the output of a Series 66 TDMS are used as the input to a series 6850 Voice-Frequency Carrier System. In turn, the information on this system provides the input to a Series 65 PLC system for transmission over a high-voltage powerline.

A photograph of a typical terminal, suitable for relay-rack mounting, is shown as Figure 4.3. In Figure 4.4 a block diagram of a typical system is shown, including the v-f carrier equipment. Details of RFL's time-division multiplexing system are given in the following:

RFL Series 66: Bulletin SBTDMS
Time-Division-Multiplex Fundamentals: Bulletin 66SYS-1
Encoders and Decoders: Bulletin SBENC.

TELEMETRY SYSTEMS

General

Telemetry systems are used to transmit the magnitude of a measured quantity to a remote location where it is displayed, used for control, or both. Modern telemetry systems are available in two forms, analog and digital.

In the former, an electrical analog of the quantity in form suitable for transmission is generated at the source. At the receiving terminal the analog signal is again converted to the magnitude of the measured quantity.

In a digital telemetry system, the quantity measured is digitized at the source and sent over the communication system as binary data. At the receiver it may be displayed digitally or used for direct control, each without further conversion. If needed, it may be restored to analog form by using a digital-to-analog converter.

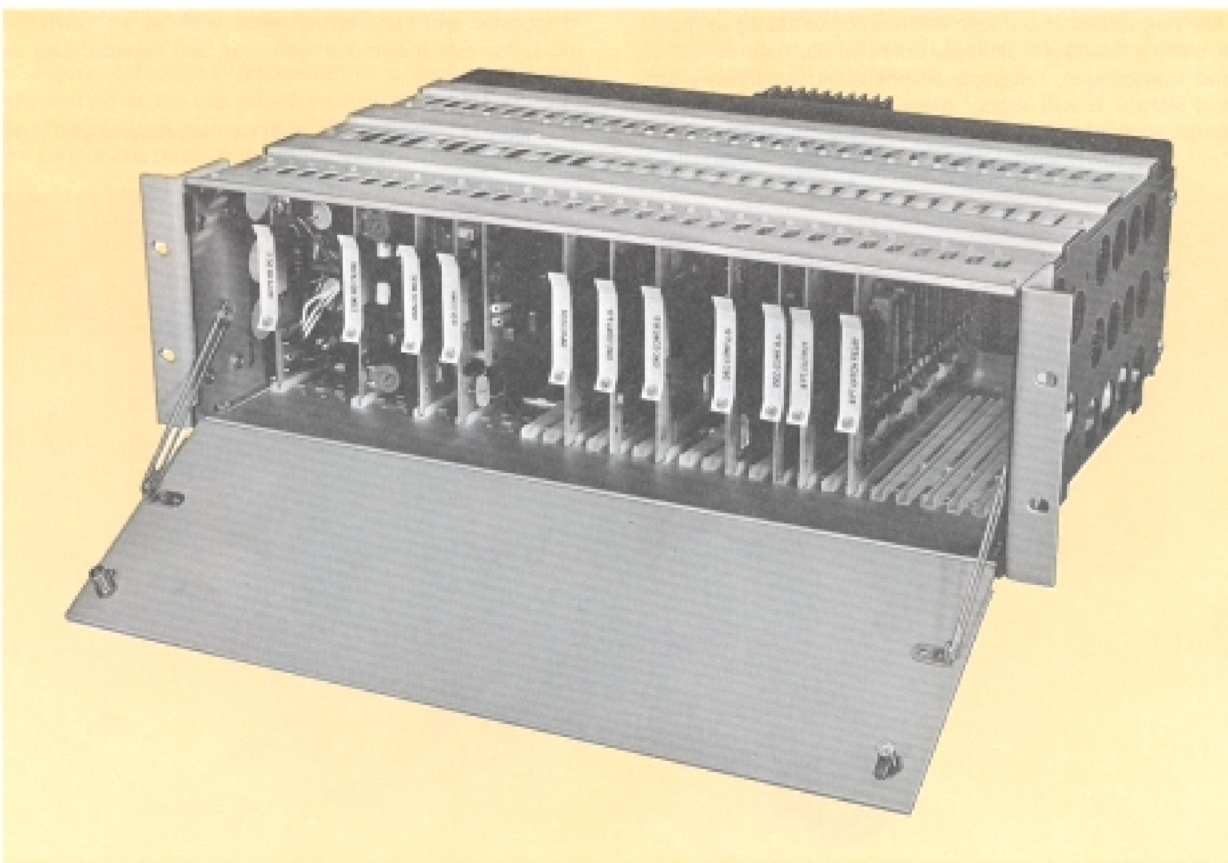


Figure 4.3. Typical supervisory-control terminal using Series 66 TDMS equipment.

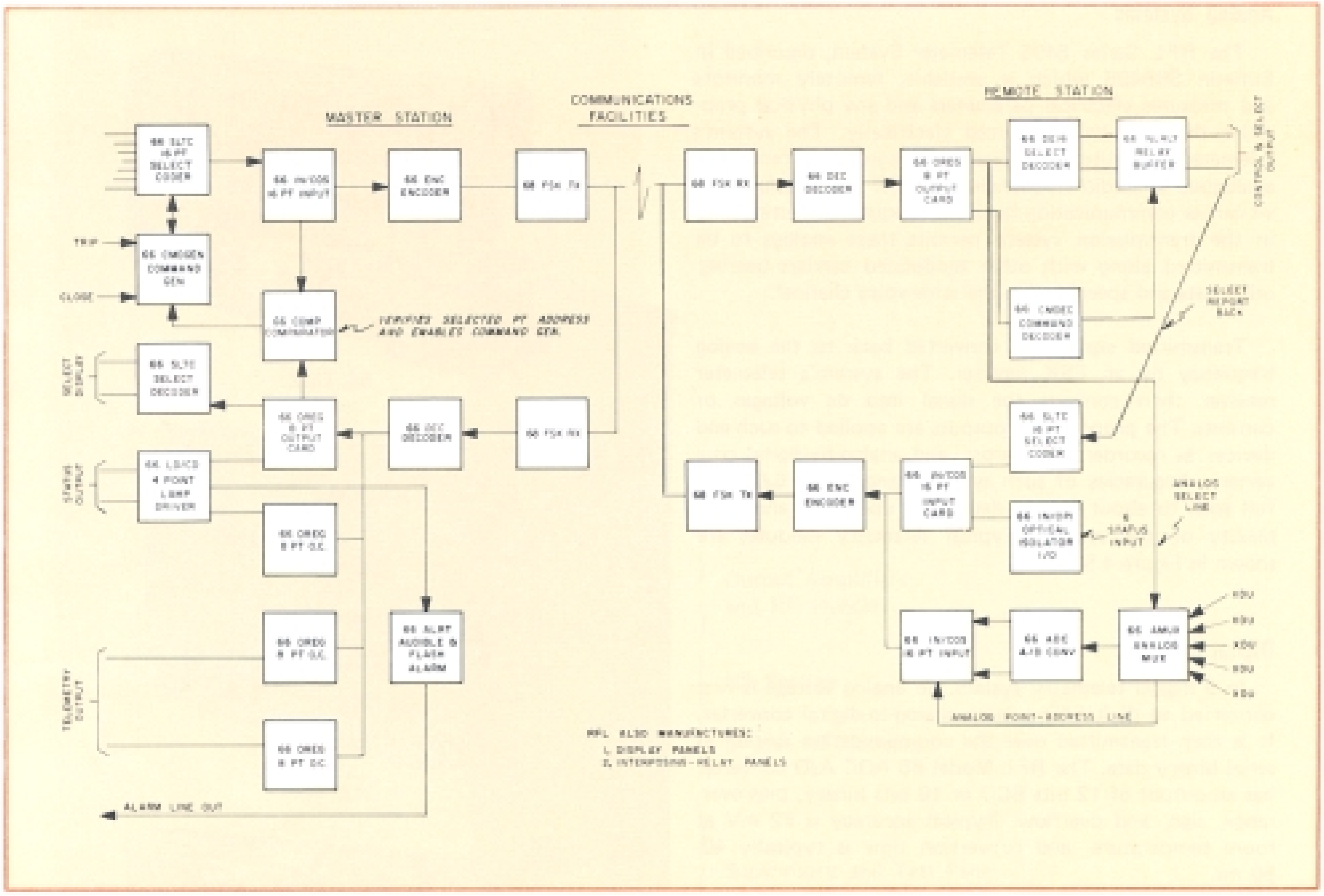


Figure 4.4. Block diagram of typical Series 66 TDMS.

Analog Systems

The RFL Series 6405 Telemeter System, described in Bulletin SB6405 which is available, remotely monitors and measures electrical parameters and any physical properties that can be transduced electrically. The system's telemeter transmitter converts dc voltages or currents into analogous subaudio frequencies for FSK transmission over v-f carrier communication facilities. Frequency multiplexing in the transmission system permits these analogs to be transmitted along with other modulated carriers bearing other data and speech, all in the same voice channel.

Transmitted signals are converted back to the analog frequency by an FSK receiver. The system's telemeter receiver then converts the signal into dc voltages or currents. The proportional outputs are applied to such end devices as recorders, indicators, and analog-to-digital converters. Accuracies of such systems range from 0.02% of full span to about 0.25%, depending upon cost and complexity of the system. Typical telemetry modules are shown in Figure 4.5.

Digital Systems

In a digital telemetry system, an analog voltage is first converted to digital form in an analog-to-digital converter. It is then transmitted over the communication system as serial binary data. The RFL Model 66 ADC A/D converter has an output of 12 bits BCD or 10 bits binary, plus overrange, sign, and overflow. Typical accuracy is ±2 mV at room temperature, and conversion time is typically 40-50 ms.

Different input signals may be sent sequentially, in groups of eight, by preceding the Model 66 ADC with a Model 66 AMUX Analog Multiplexer which selects each input signal in turn. A data sheet is available.

At the receiving terminal, digital displays with 2 or 4 digits are available for display of the telemetered quantity, or a 3-1/2-digit digital display with sign may be selected. If the analog signal from the sending point is to be reconstituted from the digital data, a Model 66 DAC Digital-to-Analog Converter is required.

Digital telemetry systems require the use of a Series 66 TDMS system as the terminal equipment. Data on digital telemetering and other elements of the 66 TDMS System are given in Bulletin SBTDMS.

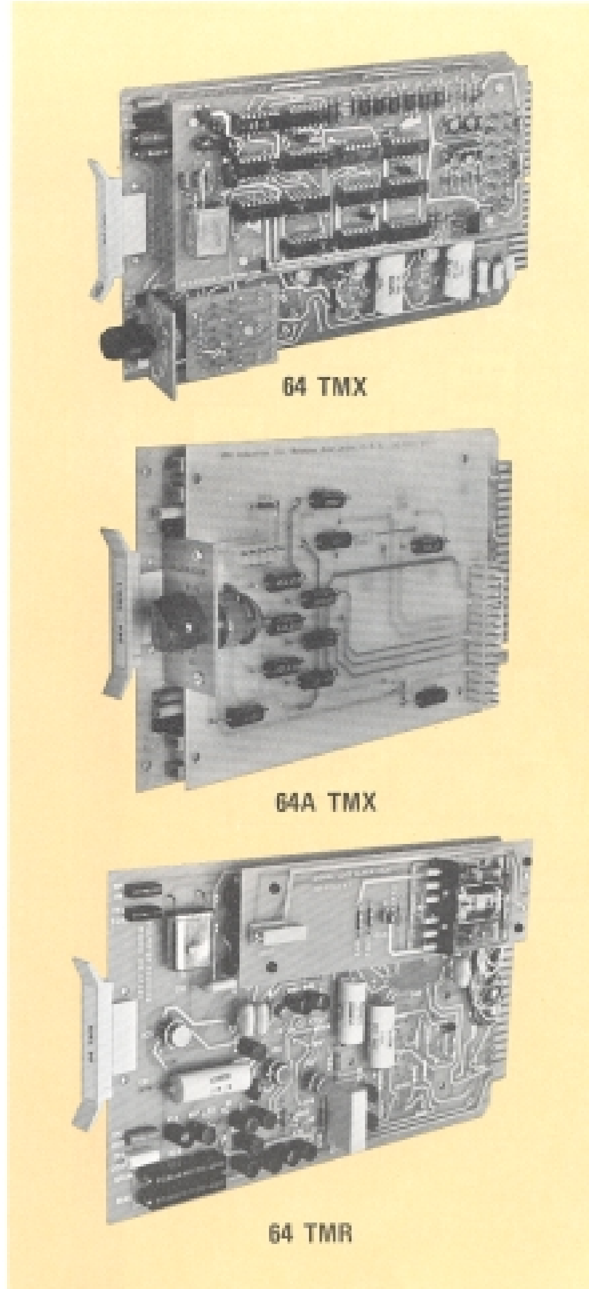


Figure 4.5. Typical telemetry modules.

Chapter 5

SPECIFICATIONS SERIES 65 PLC

OVERALL SYSTEM SPECIFICATIONS

RF SECTION

Mode of Operation
Amplitude-modulated single-sideband transmission, with suppressed or reduced carrier.

RF Range
30 to 512 kHz.

Nominal Channel Bandwidth
4 kHz standard.

Frequency Spacing
Multiples of 4 kHz are standard.

RF Impedance
50 ohms standard, others available.

Return Loss
12 dB or greater.

Frequency Stability
±10 ppm standard, ±3 ppm with optional temperature-controlled oven, or 0 ppm with synchronizer.

Power Output
40-watt PEP is standard, others available are 2 watt, 5 watt, 80 watt, 120 watt all PEP.

Harmonic Output
80 dB down or greater.

Unwanted Sidebands
80 dB down or greater.

Intermodulation
60 dB down, or greater, referred to PEP.

Receiver Sensitivity
−20 dBm at input to the downconverter.

Image Rejection
80 dB or greater.

Selectivity
100 dB or greater 4 kHz from either side of the RF band edge measured at the AF output.

IF Suppression
80 dB or greater.

AF SECTION

Bandwidth
300 to 3425 Hz standard, speech-plus systems are available.

Distortion
Equal to or less than 1% for an 800-Hz test tone at +7 dBm.

Frequency Response
See Figure 5.1 (a).

System-Level Regulation
Audio-frequency output varies no more than ±1.0 dB for rf variation of +15 to −25 dB. See Figure 5.1 (b). Can be skewed either ±5, ±10 dB, or ±15 dB.

Sending Level
Four-wire: −16 dBm.
Two-wire: 0 dBm.

Receiving Level
Four-wire: +7 dBm.
Two-wire: −4 dBm.

Impedances
Four-wire send: 600 ohms balanced.
Four-wire receive: 600 ohms balanced.
Two-wire: 600 or 900 ohms balanced.

Return Loss
26 dB or greater.

Longitudinal Balance
60 dB or greater.

Idle Noise
−60 dBmop or less.

Surge Capability

The system meets the SWC requirements of ANSI C37.90a-1974, IEEE Std. 472-1974, for line and power connections all others meet IEC-495.

Operating Temperature Range

In the ambient temperature range from 0 to 45°C the equipment meets all its specifications. From −20 to 60°C, operation is reliable.

Primary Power

Choice of 24, 48, or 120 Vdc, −10, +20%. Systems with two-watt and five-watt terminals can be operated also from 115/230 Vac ±13%, 48-65 Hz, and from 12 Vdc, −10, +20%, (other input requirements can be accommodated.)

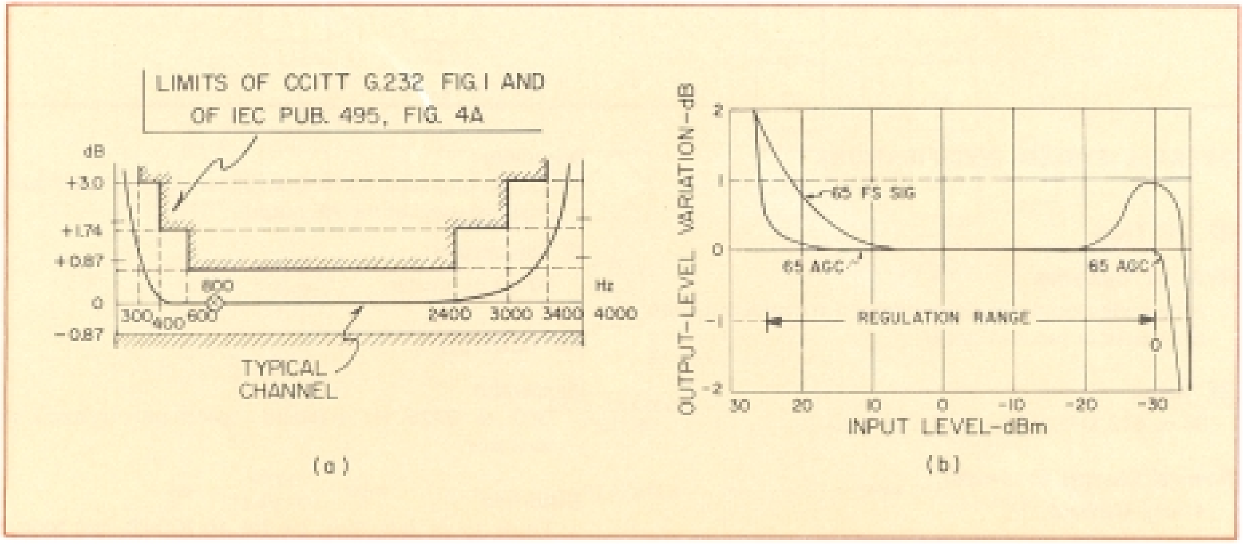


Figure 5.1. Curves of system characteristics. At (a) audio-frequency response of a typical channel is shown. At (b) system-regulation curves are given with the system under control of either a Model 65 FS SIG with optional AGC circuit or under control of the Model 65 AGC.

The RFL Series 65 PLC System meets all pertinent IEC and CCITT specifications.

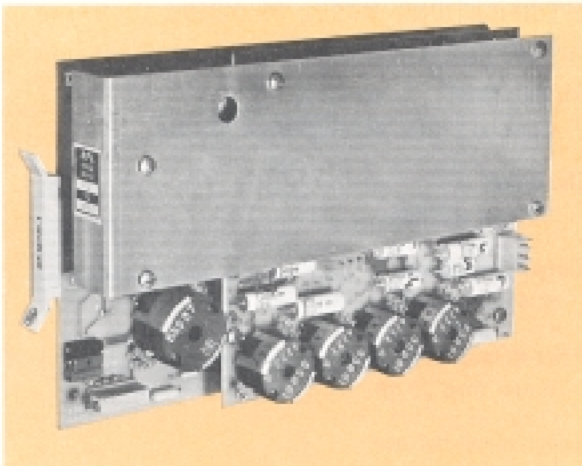


Figure 5.2. Model 65 MOD. The Model 65 DEMOD is similar in appearance.

MODEL 65 MOD SSB MODULATOR

The Model 65 MOD Modulator card is the a-f-input port of the Series 65 SSB System. Its input is audio information in a 4-kHz band, which is modulated and frequency translated by the Model 65 MOD to any 4-kHz-wide channel in the range from 4 to 60 kHz.

The block diagram, Figure 5.3, outlines the design philosophy for the modulator and its application in the System. A low-pass filter for speech-frequency signals, and a bandpass channel filter for the high-frequency output signal, both plug on to the basic circuit card. The speech filter and channel filter may be located external to the equipment, or they may be bypassed, according to the application. The channel filter's frequency range is selected for the application. The carrier signal is generated locally, and the crystal frequency must be chosen for the desired carrier frequency.

The Model 65 MOD-1 provides all the functions described here and on the block diagram. The Model 65 MOD-2 adds a constant-temperature oven for the quartz crystal to improve its frequency stability.

SPECIFICATIONS, MODEL 65 MOD

AF CHARACTERISTICS

Input Level

The range is 0 to −20 dBm. Nominal input level is −16 dBm.

Input Impedance

600 ohms balanced.

Echo-Return Loss

Better than 26 dB.

Longitudinal Balance

Better than 60 dB.

Limiting Level

Limiting will occur at a level 4 dB above the nominal setting.

RF CHARACTERISTICS

Output Level

Nominal −20 dBm.

Termination Impedance

75 ohms.

Modulator-Carrier Leak

−45 dBmo (0 dBmo = −20 dBm).

Out-of-Band Intermodulation Products

−60 dBmo (0 dBmo = −20 dBm).

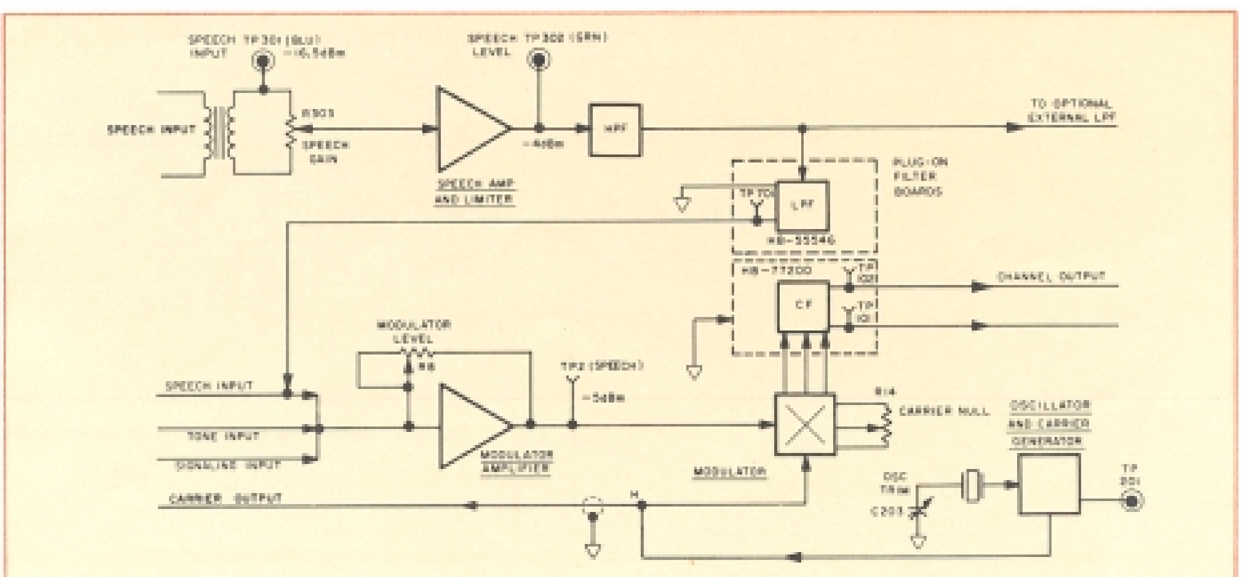


Figure 5.3. Block diagram of Model 65 Modulator.

MODEL 65 DEMOD SSB DEMODULATOR

The Model 65 DEMOD Demodulator card is the a-f-output port of the Series 6515 SSB System. Its input is a single-sideband, high-frequency signal in any of the 4-kHz-wide bands in the range from 4 to 60 kHz. The input signal is demodulated to audio baseband, the output signal of the device. Optional automatic-gain control (AGC) provides for holding the level of the speech-output signal to a predetermined value.

The block diagram, Figure 5.4, outlines the design philosophy used, and the application of the demodulator in the System. The speech filter and channel filter may be located external to the equipment, or they may be bypassed, according to the application. The plug-on bandpass channel filter's frequency range is selected for the application. The card accepts a carrier-frequency signal from an external source or, in special cases, a local oscillator may be provided directly on the card.

Model 65 DEMOD-1 provides all the functions described here excepting carrier generation and AGC. Model 65 DEMOD-2 adds AGC to the foregoing.

SPECIFICATIONS, MODEL 65 DEMOD

AF CHARACTERISTICS

Output Level

Nominal 7 dBm, limiting level is 12 dBm.

Distortion

1% max at 7 dBm.

Output Impedance

600 ohms balanced.

Echo-Return Loss

Better than 26 dB.

Longitudinal Balance

Better than 60 dB.

RF CHARACTERISTICS

Input Level

−20 dBm.

Driving Impedance

75 ohms.

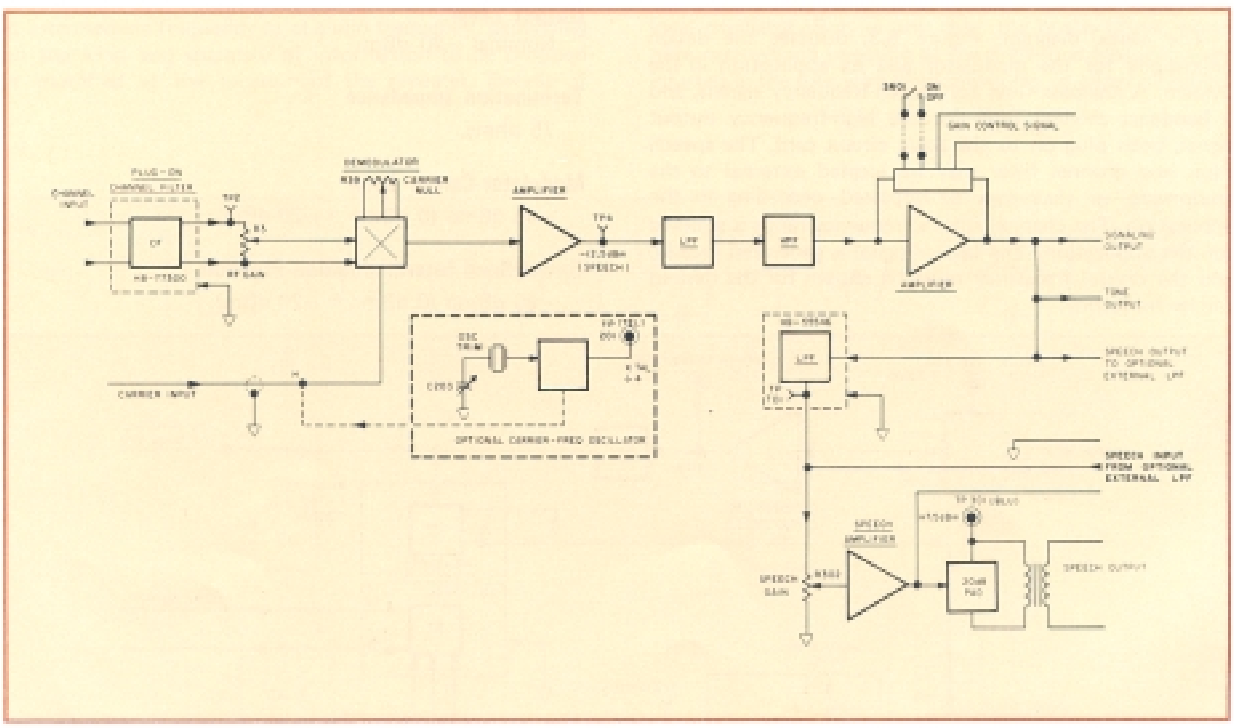


Figure 5.4. Block diagram of Model 65 Demodulator.

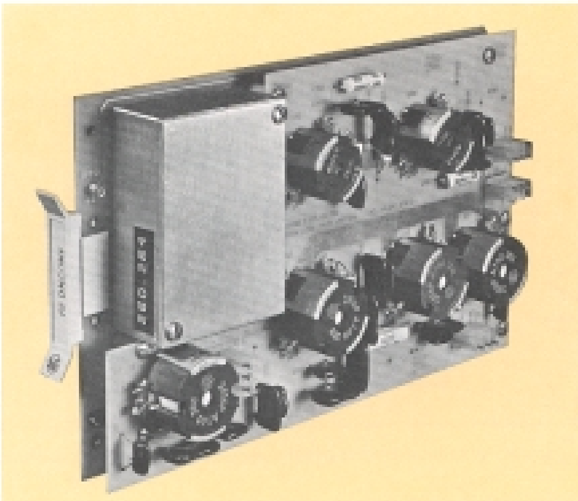


Figure 5.5. Model 65 DNCONV. Model 65 UPCONV is similar in appearance.

MODEL 65 UPCONV UPCONVERTER, and MODEL 65 DNCONV DOWNCONVERTER

These two modules provide the frequency translation necessary to extend the band of communication frequencies used on the powerline above the upper limit of 60 kHz, the higher frequency of the Models 65 MOD and 65 DEMOD. They extend the operating range to 512 kHz. The modules are similar in both function and circuit detail. A typical module is shown on Figure 5.5. Block diagrams appear as Figures 5.6 and 5.7.

MODEL 65 UPCONV

The input to the Model 65 UPCONV may be any signal in the frequency range from 4 to 60 kHz, usually one or more 4-kHz-wide channels of modulated audio baseband from the RFL Model 65 MOD Modulator.

The upconverter translates the frequency of its input signals to the frequency range between 60 and 512 kHz; and its output circuit is designed to interface with a power-line-carrier amplifier, or with microwave baseband equipment.

The upconverter uses a carrier generator which depends upon a crystal oscillator operating in the range from 4 to 9 MHz. Stability of the frequency generated is improved by addition of an optional oven which will hold the crystal at 60 ±3°C.

Output filters for the upconverter are plugged on the basic circuit card and are designed for the frequency of operation of each individual system.

MODEL 65 DNCONV

The input to the Model 65 DNCONV may be any signal in the frequency range from 60 to 512 kHz. Such signals usually originate from powerline-carrier or microwave transmission equipment, and they frequently carry several frequency-multiplexed channels.

The input signals to the downconverter are translated to the frequency range between 4 and 60 kHz. In the Series 6515 System, this output signal will consist of one or more carrier-frequency channels.

The downconverter uses a carrier generator which depends upon a crystal oscillator operating in the range from 4 to 9 MHz. Stability of the frequency generated is improved by addition of an optional oven which will hold the crystal at 60 ±3°C.

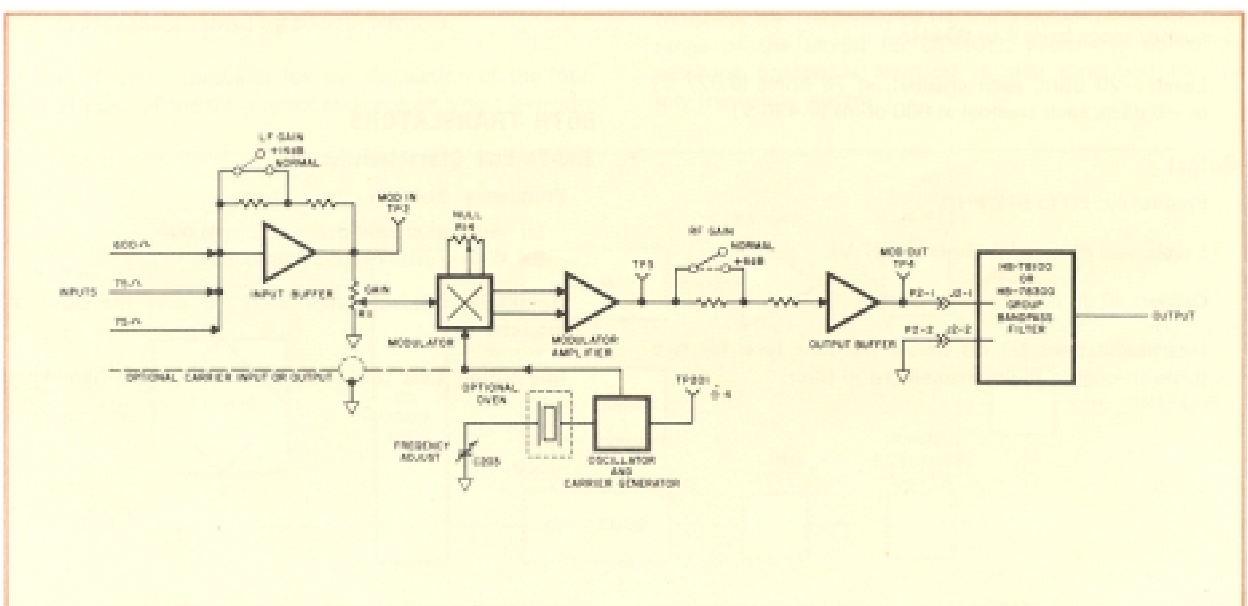


Figure 5.6. Block diagram of Model 65 UPCONV.

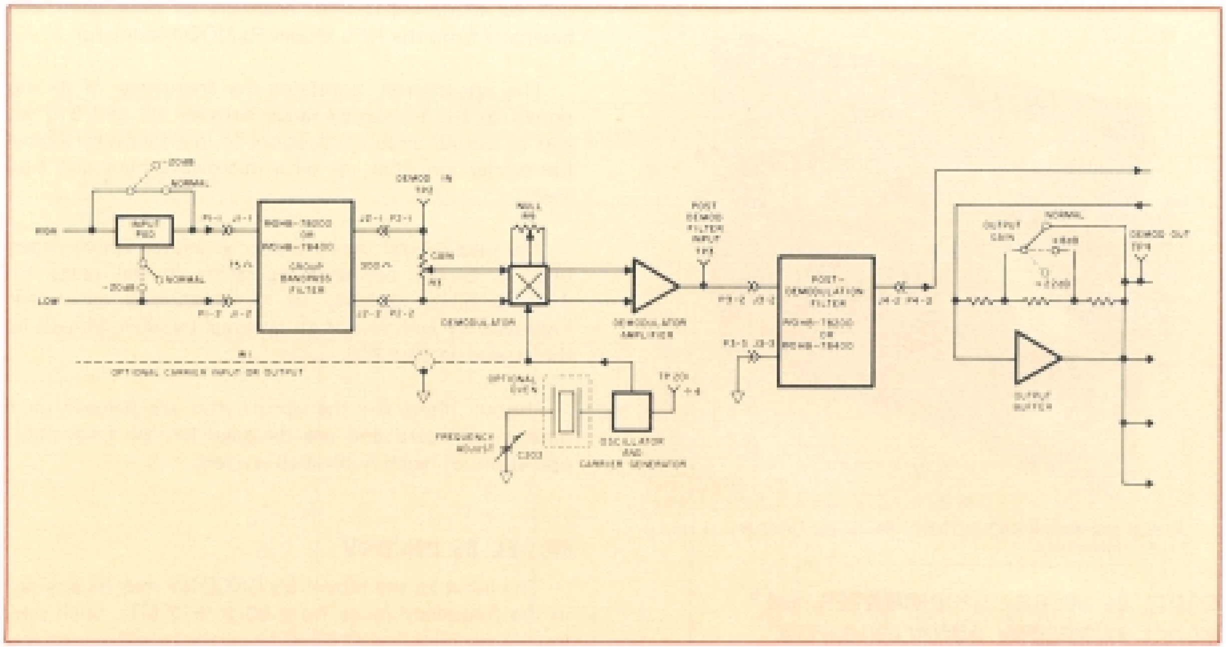


Figure 5.7. Block diagram of Model 65 DNCONV.

Output filters are plugged on the basic circuit card and are designed for the frequency of operation of each individual system.

The post-demodulation section of the filters for the down-converter may be designed for different cutoff frequencies.

SPECIFICATIONS

MODEL 65 UPCONVERTER

Input

Frequency: SSB channels of 4-kHz bandwidth in the frequency range from 4 to 60 kHz.

Level: −20 dBm, each channel, at 75 ohms (0.027 V) or −5 dBm, each channel at 600 ohms (0.436 V).

Output

Frequency: 60 to 512 kHz.

Level: −20 dBm at 75 ohms (0.027 V).

Carrier: 40 dB below test-tone level.

Intermodulation: 60 dB below test-tone level for two tones through a multi-channel group filter.

MODEL 65 DOWNCONVERTER

Input

Frequency: 60 to 512 kHz.

Level: −20 dBm through 75 ohms (0.027 V).

Output

Frequency: SSB channels of 4-kHz bandwidth in the frequency range from 4 to 60 kHz.

Level: −20 dBm (0.027 V at 75 ohms or 0.077 V at 600 ohms).

BOTH TRANSLATORS

End-To-End Characteristics

Frequency Stability:

(a) Without crystal oven: ±30 ppm max.
(b) With crystal oven: ±6 ppm max.

Amplitude Characteristic: ±0.4 dB over the band of interest.

Idle Noise: Less than −60 dBmo for 4-kHz bandwidth.

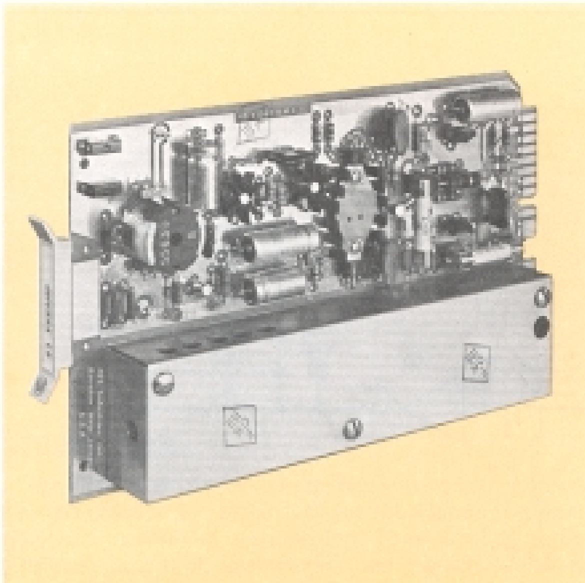


Figure 5.8. Model 65 PREAMP Preamplifier.

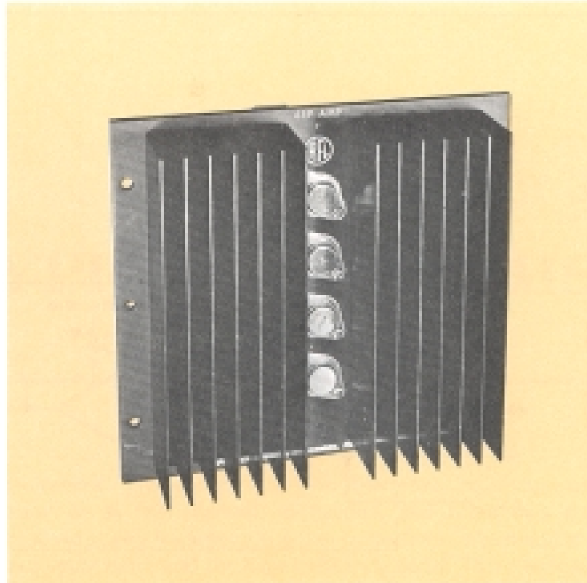


Figure 5.9. Model 65 PWRAMP 40-Watt Power Amplifier.

MODEL 65 PREAMP PREAMPLIFIER, and MODEL 65 PWRAMP POWER AMPLIFIER

The Models 65 PREAMP and 65 PWRAMP are elements of the Series 65 Single-Sideband Carrier System. The Model 65 PREAMP will supply the driving power for up to six units of the Model 65 PWRAMP. These may be paralleled for output power up to 240 watts, selectable in steps of 40 watts. When paralleled, gains and phases are equalized so that maximum power is delivered to the common, 50-ohm load.

The Model 65 PWRAMP must be driven by the Model 65 PREAMP. It cannot be driven directly from either the Model 65 MOD Modulator or the Model 65 UPCONV Upconverter. The Model 65 PREAMP is not used as an output amplifier. For output power at lower levels, the Model 65 AMP Two-Watt Amplifier, is recommended.

The Model 65 PREAMP is constructed on a circuit card which plugs into a Model 68 Chassis, like all other circuit modules of the Series 65 System. Because of its greater size, however, the power amplifier is mounted independently on the relay rack.

Figure 5.8 shows the appearance of the Model 65 PREAMP. The Model 65 PWRAMP is shown as Figure 5.9.

SPECIFICATIONS

MODEL 65 PREAMP

Frequency Response

Two ranges, either 4-60 kHz or 30-512 kHz, each ±0.5 dB with respect to midband response.

Harmonic Distortion

400 kHz or less: −55 dBmo or better.
500 kHz or less: −50 dBmo or better.

Third-Order Intermodulation Distortion

−60 dBmo.

MODEL 65 PWRAMP

Power Output

Up to 40 watts to a 50-ohm load. Amplifiers may be paralleled for greater power.

Frequency Response

Two ranges, either 4-60 kHz or 30-512 kHz, each ±1 dB with respect to midband response.

Harmonic Distortion

Second harmonic is 40 dB or better below fundamental.

Intermodulation Distortion

400 kHz or less: −60 dB or better.
500 kHz or less: −56 dB or better.

Return Loss

10 dB or greater.

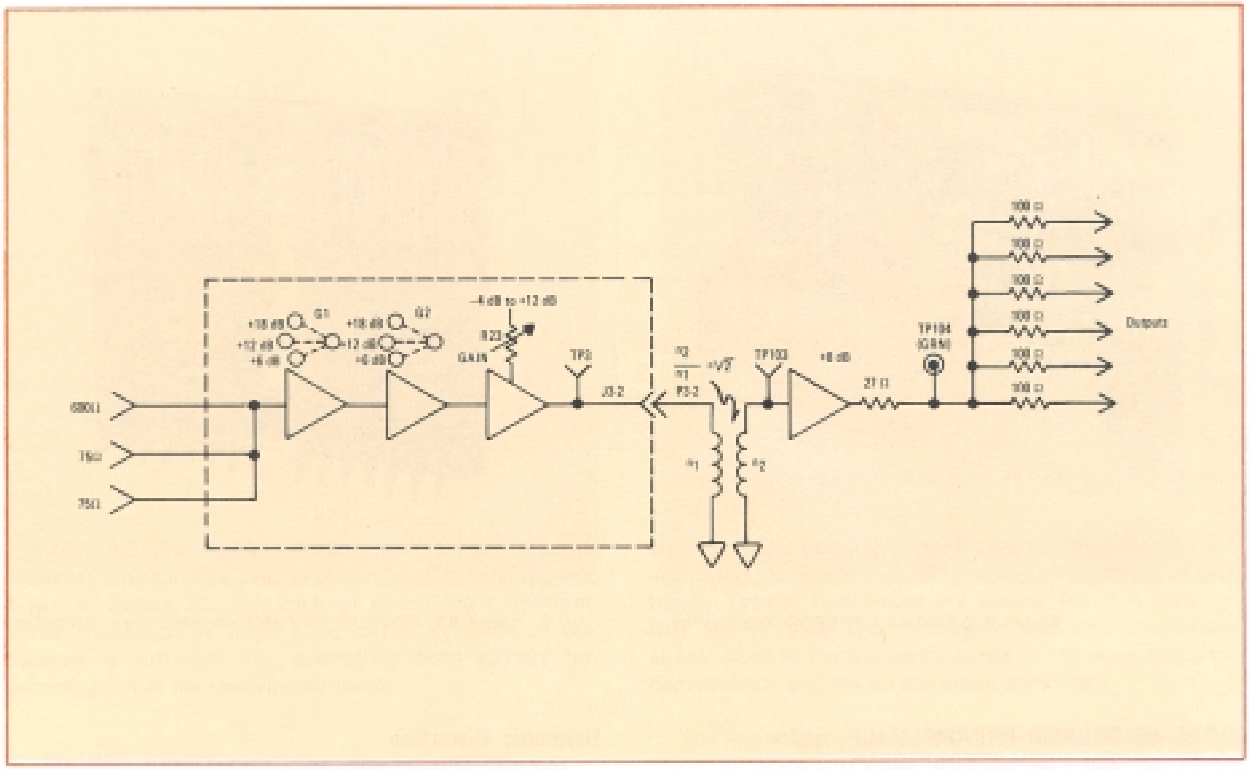


Figure 5.10. Block diagram of circuit, Model 65 PREAMP.

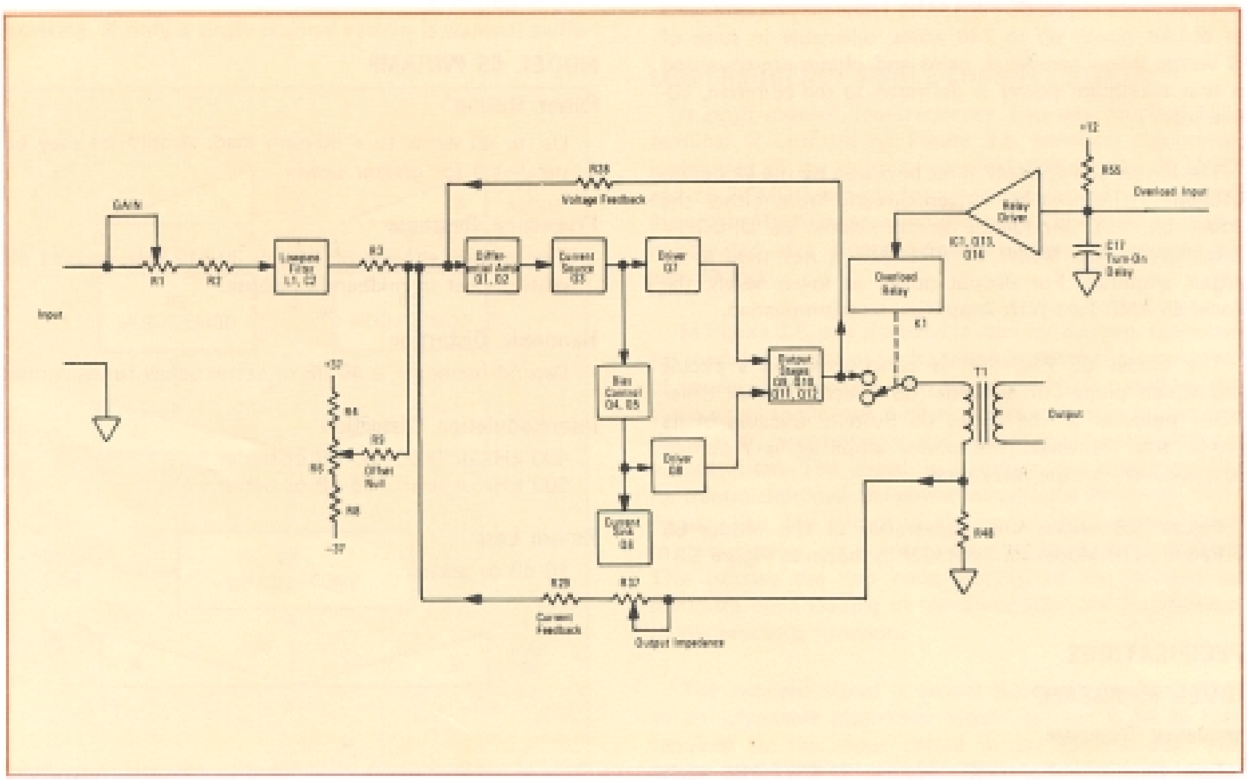


Figure 5.11. Block diagram of circuits, Model 65 PWRAMP.

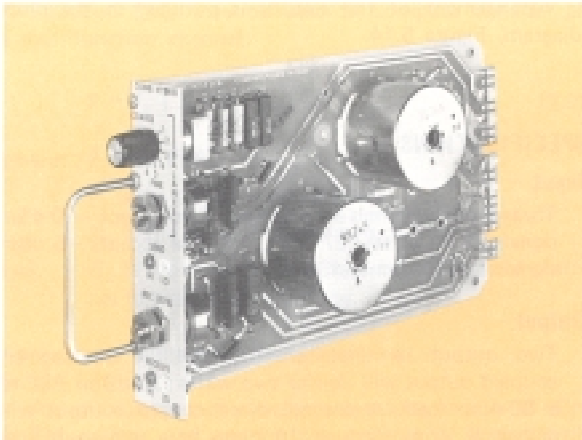


Figure 5.12. Model 65 COMB HYB.

MODEL 65 COMB HYB COMBINATION HYBRID

The Model COMB HYB, shown in Figure 5.12, is a skewed RF hybrid which is connected between a two-wire transmission medium and the single-sideband terminal equipment to perform the following functions.

(a) The rf hybrid provides isolation between the receiving and transmitting sections of the terminal, and it provides for two-wire coupling to the line-tuning unit.

(b) It contains an attenuator to control the level of received signal at the input to the receiving system.

The Model 65 COMB HYB presents a fixed input impedance of 50 ohms or 25 ohms to the transmitting amplifier or filter with which it is used, and a source impedance of approximately 75 ohms to the input of the receiving system. The output of the hybrid, which feeds the line-tuning unit, is set at the factory to match the impedance of the system with which it is to be used.

TECHNICAL SUMMARY

HYBRID SECTION

Insertion Loss

Transmitting: Approximately 0.3 dB, 60 to 500 kHz.
Receiving: Approximately 12.5 dB, 60 to 500 kHz.

Transhybrid Loss:

–40 dB, or better, 60–500 kHz, when operating with specified termination impedance.

Internal-Balance Range

0.5 to 1.95 times the specified two-wire impedance.

SYSTEM SPECIFICATIONS (System plus 65 COMB HYB)

Power Capacity

40 watt rms.

Two-Wire-Line Impedance

50 or 75 ohms are standard. Others available.

Second-Harmonic Distortion

80 dB, or better, below-fundamental

Third-Order Intermodulation Products

60 dB, or better, below fundamental

Surge Protection

The system meets the SWC requirements of ANSI C37.90a–1974, IEEE Std. 472–1974.

Operating-Temperature Range

In the ambient temperature range from 0 to 45°C the equipment meets all its specifications. From –20 to 60°C, operation is reliable.

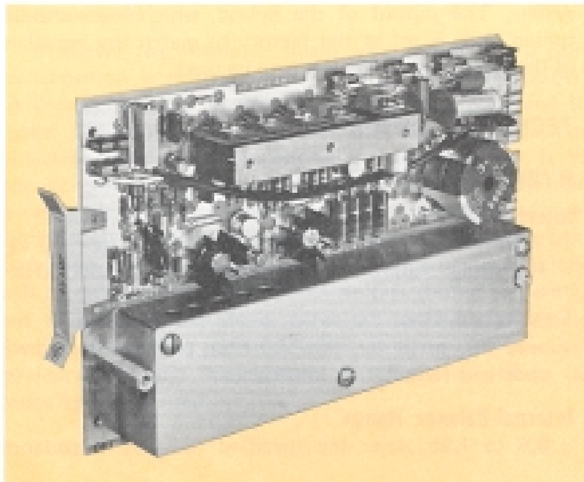


Figure 5.13. Model 65 AMP Two-Watt Amplifier.

MODEL 65 AMP TWO-WATT AMPLIFIER

The Model 65 AMP provides for raising the level of the carrier generated for PLC transmission to the sending level necessary for the terminal, providing the required level is two watts or less.

Input signals are accepted at source-impedance levels of either 75 or 600 ohms. Voltage gain is adjustable, by changing location of jumpers, in steps of 6 dB, from 12 to 52 dB. A continuous gain control, for adjustment between steps, has a range of 16 dB. Maximum power output is 2 watts (33 dBm).

The output signal is transformer-coupled to a 50-ohm load. Either of two output transformers is available. Option HB-55598 provides a transformer with a frequency range of 30 to 500 kHz. Transformer Option HB-55599 operates over the frequency range from 4 to 60 kHz.

Direct-coupled output, through a 10-ohm isolating resistor, from the emitters of the complementary output pair, is provided for driving a higher power amplifier, if needed.

Arrangement of the circuits is outlined in the block diagram, Figure 5.14.

SPECIFICATIONS

Input

Three input terminals are provided: Two accept –20 dBm signals at 75 ohms (27 mV), and one accepts –5 dBm signals at 600 ohms (436 mV).

Output

Two output terminals are provided: A transformer-coupled output will deliver two watts (33 dBm) PEP to a 50-ohm load. A second, direct-coupled output will deliver 5 Vrms through 10 ohms to a minimum load of 100 ohms.

Voltage Gain

Maximum voltage gain is 52 dB, adjustable from 12 to 52 dB in 6-dB increments augmented with a continuous gain-control range of 16 dB. Gain of stages 1 and 2 is set with jumpers at 6, 12, or 18 dB. Stage 3 is adjustable from –4 to 12 dB, and Stage 4 has a fixed gain of 4 dB.

Frequency Response

4 to 500 kHz ±1.0 dB with respect to midband of two different frequency ranges: 4-60 kHz or 30-500 kHz, depending on choice of output transformer.

Harmonic Distortion

Below 60 kHz	–60 dBmo, or better
Below 100 kHz	–55 dBmo, or better
Below 400 kHz	–45 dBmo, or better
Below 500 kHz	–40 dBmo, or better

Third-Order Intermodulation Distortion

–60 dBmo, or better.

Idle Noise

–65 dBmo, or better (3 kHz bandwidth).

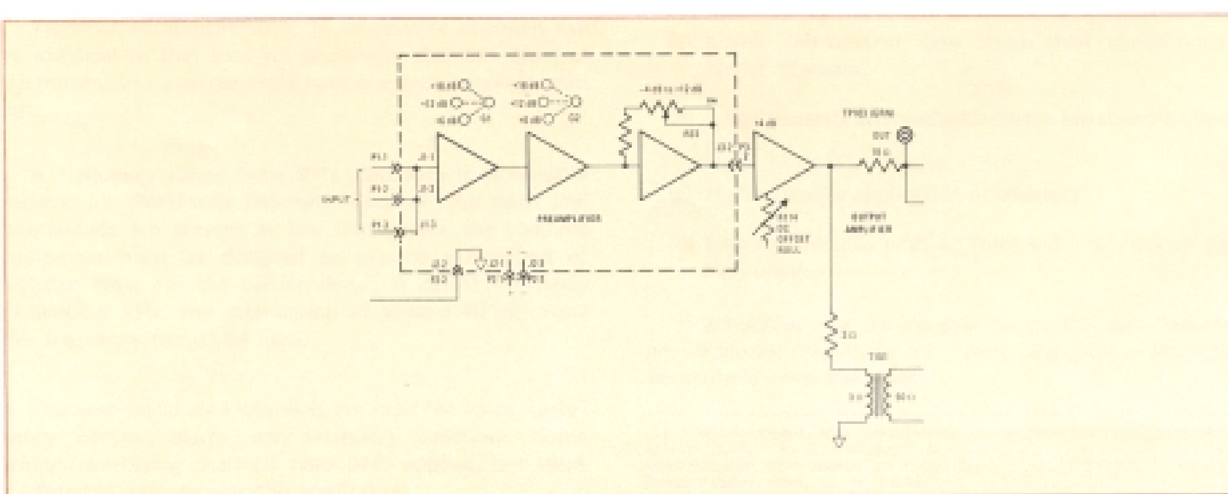


Figure 5.14. Block diagram of circuits, Model 65 AMP.

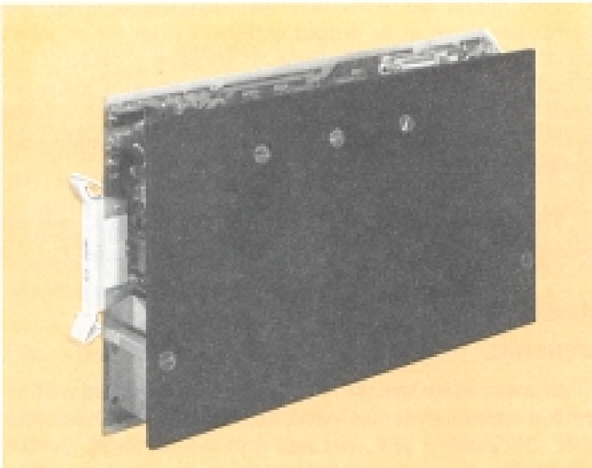


Figure 5.15. Model 65 AMP Five-Watt Amplifier.

MODEL 65 AMP FIVE-WATT AMPLIFIER

The Model 65 AMP is an element of the RFL Series 6515 Single-Sideband Carrier System. This five-watt amplifier is used to raise the output level of the Model 65 UPCONV to the sending level necessary for the carrier terminal.

Input signals are accepted at source-impedance levels of either 75 or 600 ohms. Voltage gain is adjustable, by changing the location of jumpers, in steps of 6 dB, from 12 to 55.5 dB. A continuous gain control, for adjustment between steps, has a range of 23 dB. Maximum power output is 5 watts (37 dBm).

The output signal is transformer-coupled to a 50-ohm load.

TECHNICAL SUMMARY

Input

Three input terminals are provided: Two accept -20 dBm signals at 75 ohms (27 mV), and one accepts -5 dBm signals at 600 ohms (436 mV). NOTE — Preamp may be jumpered when used in a system having 2 V input signals.

Output

A transformer coupled output will deliver five watts (37 dBm) PEP to a 50-ohm load.

Voltage Gain

Maximum voltage gain is 55.5 dB, adjustable from 12 to 55.5 dB in 6-dB increments augmented with a continuous gain-control range of 23 dB. Gain of stages 1 and 2 is set with jumpers at 6, 12, or 18 dB. Stage 3 is adjustable from -4 to 10 dB. Stage 4 has a gain of 9.5 dB.

Frequency Response

4 to 500 kHz ±1.0 dB with respect to midband of two different frequency ranges: 4-60 kHz or 30-500 kHz, depending on choice of output transformer.

Third-Order Intermodulation Distortion:

60-350 kHz: -60 dBmo, or better
Below 500 kHz: -50 dBmo, or better

Idle Noise

-65 dBmo, or better (3 kHz bandwidth).

Power Consumption

From both plus and minus 12-Vdc power sources, the module's current drain depends upon power as follows:

Output Watts	Current Drain mA, Max.
0	350
0.5	450
1.5	700
3.0	800
5.0	900

Ambient Temperature Range

The equipment will meet all specifications in the temperature range from 0 to 45°C and it will operate in the range from -20 to 60°C.

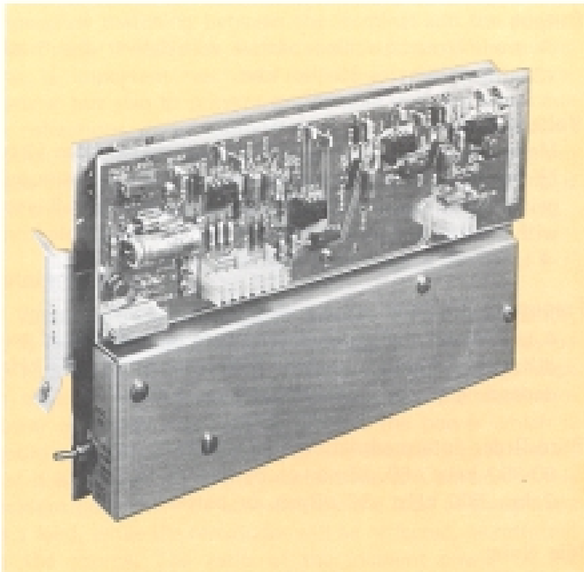


Figure 5.16. Model 65 AGC Automatic-Gain-Control Module.

MODEL 65 AGC AUTOMATIC-GAIN-CONTROL MODULE

In a typical system the Model 65 AGC is fed a pilot tone whose amplitude is compared against an internal reference. The difference between the two is used to control the gain of an amplifier in which the wideband composite signal, comprising all information passed through the system, is automatically adjusted to the nominal level. Because the pilot tone is part of the composite signal, the AGC system holds the level of all signals at the desired amplitude.

The Model 65 AGC will maintain the level of the composite signal at its output within ±1 dBm for an input variation of +15 to −25 dBmo. It contains a detector for failure of the pilot tone. Failure of that signal is indicated both by extinguishing a lamp and by release of a relay with two sets of Form-C contacts, used for additional indication. (The relay is optional).

The philosophy of the design of the 65 AGC will be evident from a study of its block diagram, Figure 5.17.

Application

GENERAL

In most cases the 65 AGC will operate in conjunction with a signalling transceiver. The pilot tone, derived from that transceiver, will be used by the 65 AGC as the regulating signal. Generally, all bandpass filtering will be done on the signalling receiver, but a separate filter card may replace the transceiver in cases where the transceiver is not used.

The pilot tone may also be derived from the Model 65 SYNC Synchronizer card.

REPEATER-STATION APPLICATION

Some applications involve translating a group of signals down to some intermediate frequency, regulating amplifying, and retranslating without reaching the baseband frequency. In such cases, the 65 AGC will be able to operate in conjunction with the Model 65 DNCONV Downconverter which would contain appropriate filters. This will enable the 65 AGC to operate from a tone located anywhere in the frequency range from 4 to 60 kHz. The 65 DNCONV will translate a tone down to baseband where it will be filtered with a selective bandpass filter. The tone will then enter the pilot-tone input of the 65 AGC.

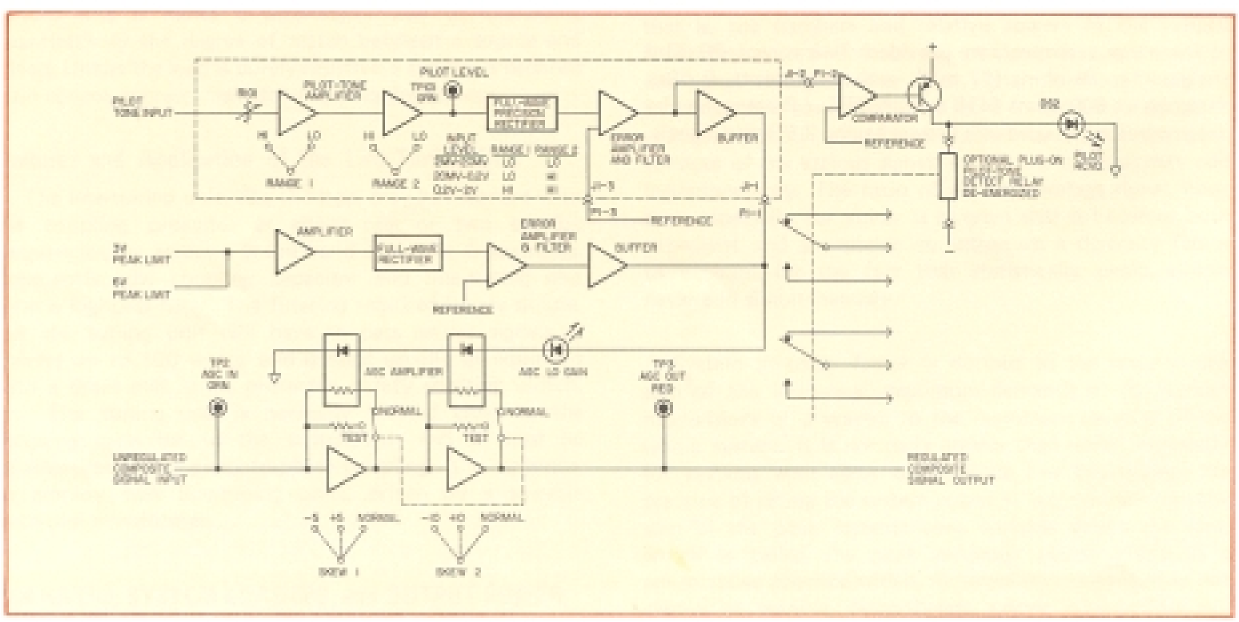


Figure 5.17. Block diagram of circuits, Model 65 AGC.

SPECIFICATIONS

Input Level

0 dBmo = 0.033 Vrms.

Dynamic Range

+15 dBmo to −25 dBmo. The dynamic range may be skewed by ±5, ±10, or ±15 dB.

Input Impedance

10K ohms, unbalanced. This reduces to 5.6K for −5 dB skew, and increases to 17.8K for +5 dB skew.

Frequency Range

4 to 60 kHz.

Output Level

0.033 Vrms.

Output Impedance

The device appears as a voltage source of zero internal impedance.

Stiffness Ratio

±1 dB change at the output for +15 to −25 dBmo change at the input.

Peak Detector and Limiter

Depending upon waveshape or on the number of discrete signals, the auxiliary peak detector will reduce the gain to maintain linear operation when the input to it peaks at 6 to 10 volts for the six-volt-limit input, or at 3 to 5 volts for the three-volt-limit input.

Distortion

Distortion and intermodulation products will be less than −60 dBmo for output levels of less than 0.33 Vrms (+20 dBmo).

Pilot-Tone Input Level

0.002 to 2.0 Vrms in three decade ranges selected by placement of a jumper.

Pilot-Tone Input Impedance

10K ohms.

Pilot-Tone Frequency

1 to 60 kHz. If the pilot tone is not band limited, an external filter must be used with the 65 AGC.

Pilot-Tone Detector

The optional pilot-tone-detect relay will drop out when the input is below −31 dBmo.

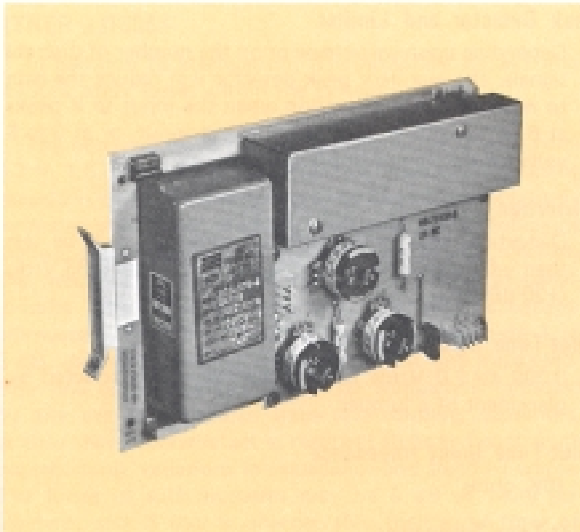


Figure 5.18. Model 65 SYNC Synchronizer module.

MODEL 65 SYNC, SYNCHRONIZER

The Model 65 SYNC eliminates frequency-translation error, by transmitting and receiving a low-level carrier signal. This signal also may be used as a pilot tone for automatic gain control.

The transmitting section takes the squarewave output of the carrier oscillator of the local Model 65 MOD modulator, shapes it to a sinusoid for transmission, and sets its proper level. The signal will normally be routed to the Model 65 UPCONV.

The input to the receiving section is the composite signal received from a remote terminal, normally from the output of the Model 65 DNCONV. The receiver filters and amplifies this signal and thereby derives the carrier-frequency reference for the Model 65 DEMOD demodulator, and also the amplitude-reference level which may be used with the Model 65 AGC.

Three plug-on filters are required. These are a transmitter filter, a receiver bandpass filter, and a receiver crystal filter. Each carrier frequency requires a different set of filters. The frequency is specified by a suffix number appended to the Model designation of the filter.

A block diagram of the circuits is shown as Figure 5.19.

SPECIFICATIONS

TRANSMITTER

Input Signal

Squarewave of carrier frequency, 12 Vp-p from a source with internal resistance of 1K ohm or less. This signal will be between 20 and 44 kHz.

Input Impedance

30K ohms minimum.

Output Signal

Sinewave at frequency of input, with harmonics at −50 dBmo, or less.

Output Level

Either 25 or 77 mV at 600 ohms. Amplitude is determined by a potentiometer and by placement of jumpers.

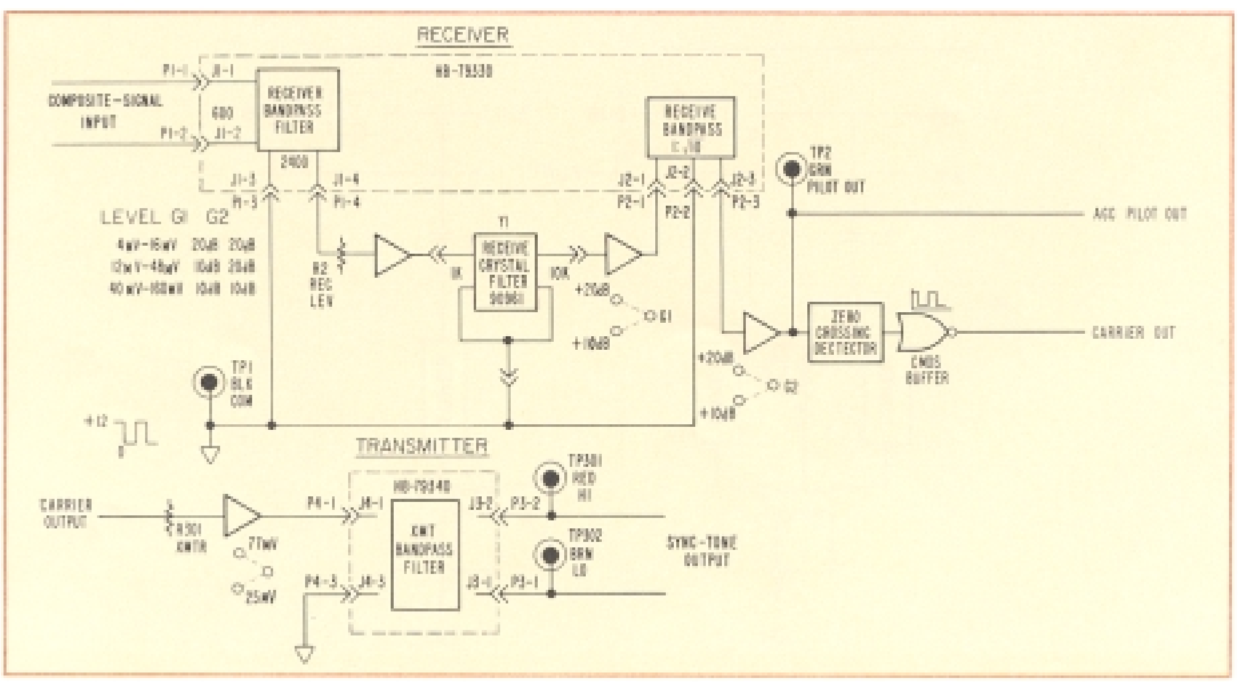


Figure 5.19. Block diagram of circuit, Model 65 SYNC.

RECEIVER

Input Signal

Composite channel signal at channel frequency, normally between 20 and 44 kHz.

Input Level

4 to 160 mVrms, adjustable by selection of gain-control jumpers and with input-level potentiometer.

Input Impedance

600 ohms.

Carrier Frequency

Maximum permissible variation of carrier frequency from nominal is ±6 Hz.

Output Signals

(a) A sinewave signal at carrier frequency, amplitude 1 Vrms ±1 dB, source impedance 100 ohms, is used as a level reference for the Model 65 AGC.

(b) A squarewave at carrier frequency, 12 Vp-p, from the output of a CMOS gate. This is used as the local-oscillator signal for the Model 65 DEMOD.

Phase Jitter

Maximum 1.5^{o} p-p.

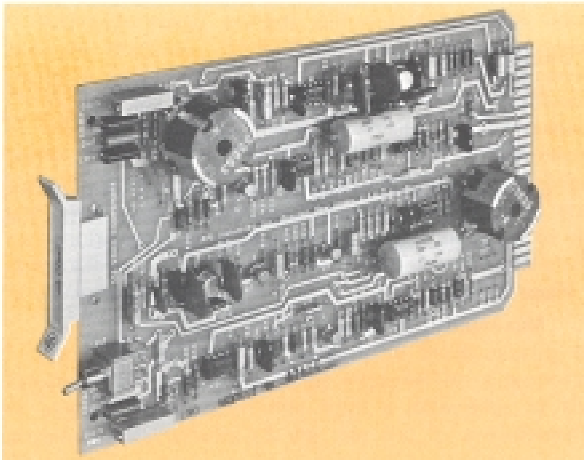


Figure 5.20. Model 65 COMP Voice Compandor.

MODEL 65 COMP VOICE COMPANDOR

The Model 65 COMP Voice Compandor is used in applications where an improvement in signal-to-noise ratio is required due to the presence of extraneous noise. The compandor can also be used together with other equipment having suitable signal levels and impedances.

Compressor and expander, together, comprise one printed-circuit-card module. The compandor complies with all conditions of CCITT Recommendation G.162.

SPECIFICATIONS

Compression Ratio: Compressor 2:1 (dB)

Expansion Ratio: Expander 1:2 (dB)

Level Stability (–25 to +60°C ±5% Voltage Variation): Compressor ±25 dB, Expander ±0.5 dB

Return Loss (300-3400 Hz): Compressor 23 dB typical, Expander 23 dB

Input Impedance: Compressor 600 Ω bal. or unbal., Expander 600 Ω unbal.

Output Impedance: Compressor 600 Ω unbal., Expander 600 Ω bal. or unbal.

Distortion: Compressor < 1% @ 0 dBmo (–16 dBm), Expander < 1% @ – dBmo (+7 dBm)

Attack-Time Limit: Compressor < 5 ms

Recovery-Time Limit: Compressor < 22.5 ms

Transient Response (Back-to-back, 12 dB/step): Compressor no observable overshoot.

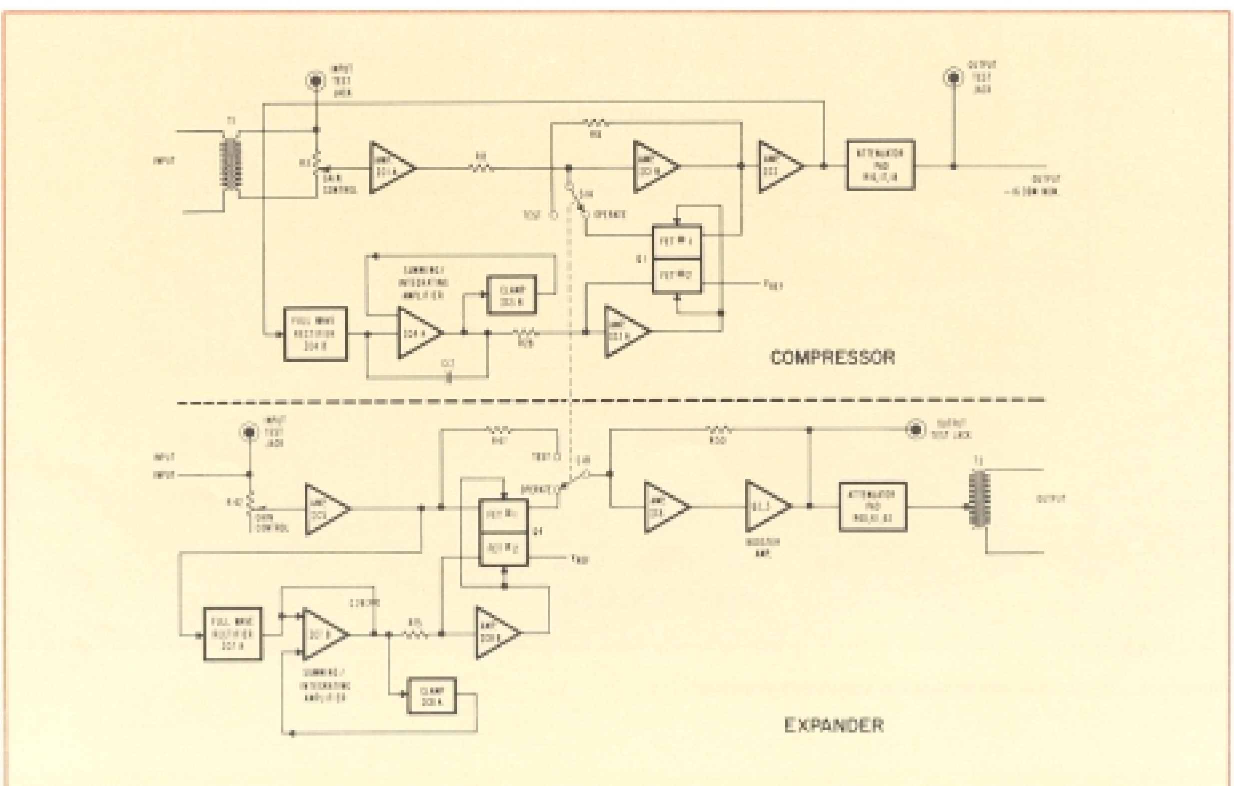


Figure 5.21. Block diagram of compressor and expander sections of Model 65 COMP.

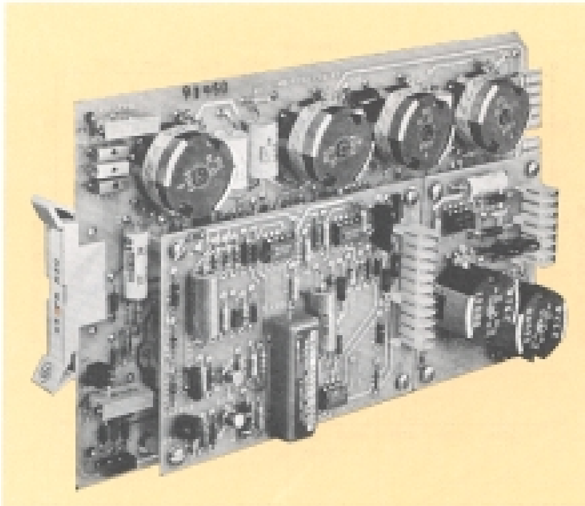


Figure 5.22. Model 65 FS SIG Frequency-Shift Signaling-Tone Transceiver.

MODEL 65 FS SIG FREQUENCY-SHIFT SIGNALING-TONE TRANSCEIVER

The Model 65 FS SIG Frequency-Shift Signaling-Tone Transceiver consists of three assemblies: the main board with the transmitter and receiver, an optional plug-on two-coil output filter, and an optional plug-on automatic-gain-control and carrier-detector card. A block diagram is shown as Figure 5.23.

The transmitter uses a state-variable oscillator, the frequency of which is shifted, by switching two resistors in and out of the circuit. The center-frequency is 3825 ±30 Hz. Both dc keying and an opto-isolator keying circuit are provided for keying the transmitter from an external source. An optional plug-on two-coil output filter is available to reduce the emission of sidebands generated when the transmitter is keyed at rates higher than twenty pulses per second.

The receiver's input uses a three-coil filter whose output drives an amplifier. The signal level can be measured at the output of this amplifier and is adjusted with a ten-turn potentiometer. Means are provided to reduce the gain, if needed. The amplifier's output feeds a Schmitt-trigger threshold detector which feeds a discriminator whose phase shift is zero at center frequency, and which provides a phase lag with an increase in frequency and a phase lead with a decrease in frequency. The output of the discriminator drives an all-pass quadrature network, the output of which, along with the output from the threshold detector, feeds an EXCL-OR gate whose duty cycle is increased or decreased by the discriminator. The output of the EXCL-OR gate feeds a lowpass filter and then a slicer which detects whether the receiver's input signal is a high or a low frequency. The presence of a high-frequency signal both lights a lamp and energizes a Form-C relay in response to the input signal.

An optional plug-on AGC and carrier-detector card uses two full-wave rectifiers, one to detect the carrier and the other to control the AGC. The dc AGC output may be used as the group AGC for the demodulator of the multiplexed single-sideband carrier system. It is accomplished by comparing the output of the full-wave rectifier with a reference. The difference drives the AGC-control circuit in the demodulator with a control current.

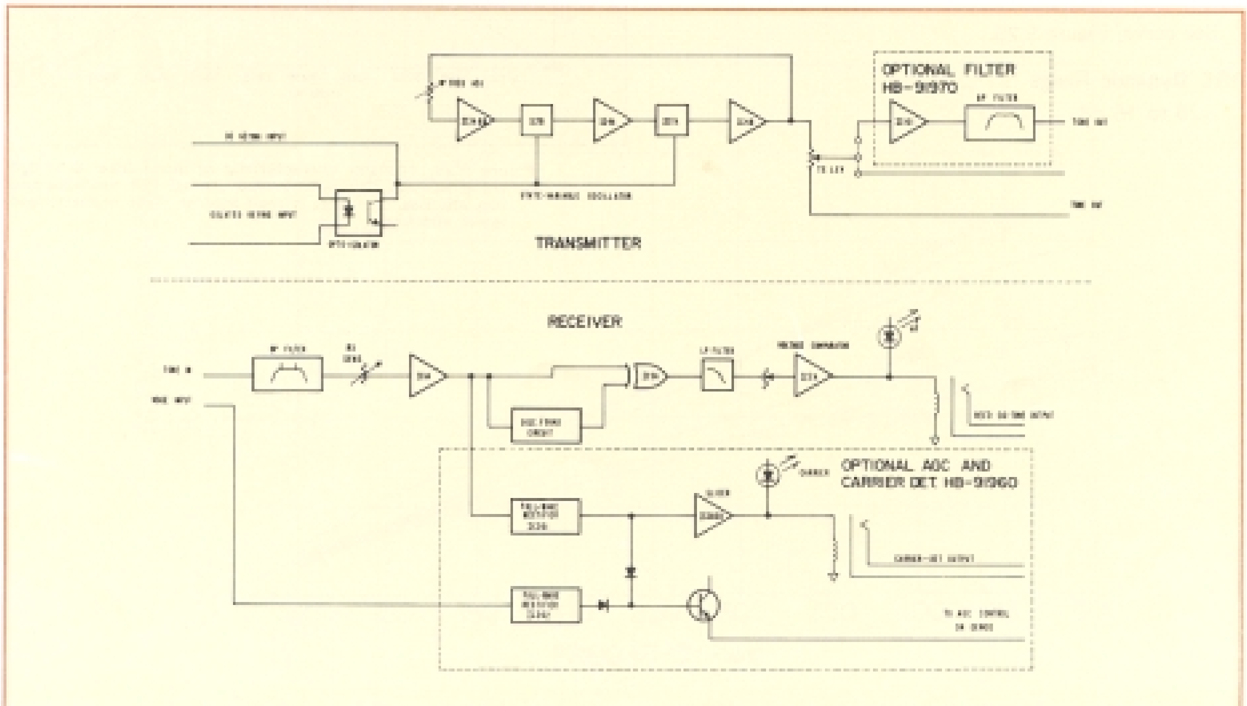


Figure 5.23. Block diagram of circuit, Model 65 FS SIG.

SPECIFICATIONS

TRANSMITTER

Frequency

3825 Hz, ±30 Hz shift.

Stability

±3 Hz over the operating-temperature range.

Harmonic Distortion

Less than 1%.

Output Level

Nominally −20 dBm.

Output-Filter Response

See curve, Figure 5.24.

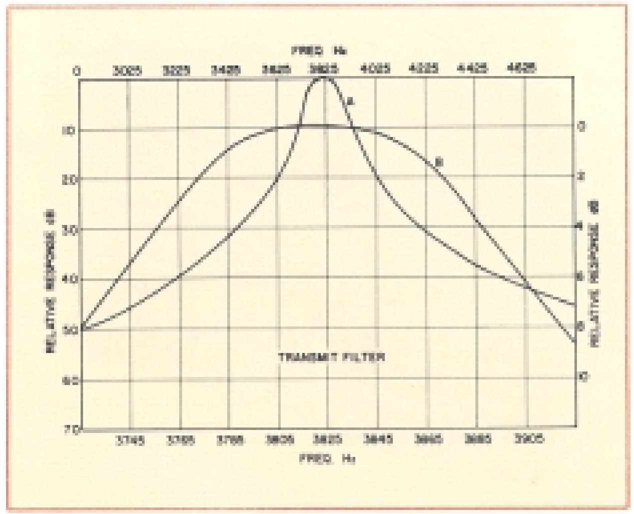


Figure 5.24. Transfer characteristic of output filter HB-91970. Curve A is read against left ordinate and top abscissa. Curve B is read against right ordinate and lower abscissa.

RECEIVER

Frequency

3825 ± 30 Hz.

Sensitivity

−35 dBm.

Input Impedance

600 ohms, balanced.

Input-Filter Response

See curve, Figure 5.25.

AGC Dynamic Range

−25 to 15 dB.

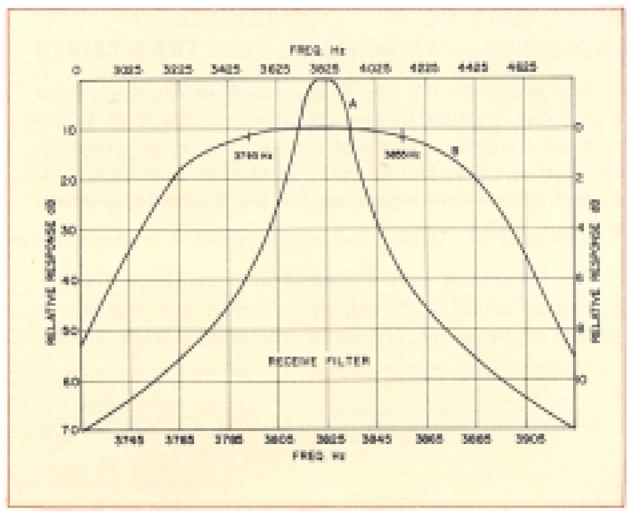


Figure 5.25. Transfer characteristic of input filter provided with receiver. Curve A is read against left ordinate and top abscissa. Curve B is read against right ordinate and lower abscissa.

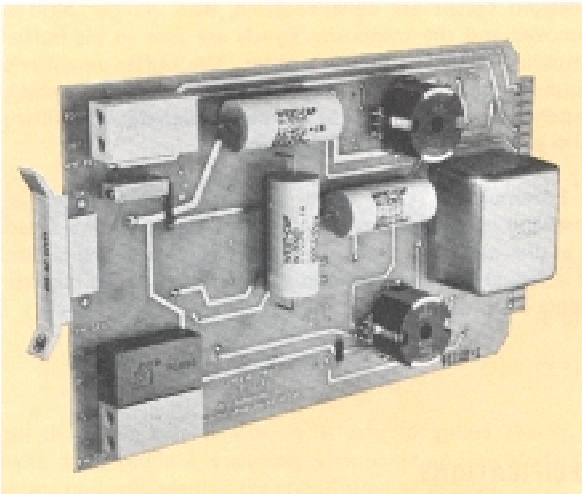


Figure 5.26. Model 65 AF HYB Audio-frequency Hybrid.

MODEL 65 AF HYB AUDIO-FREQUENCY HYBRID

The Model 65 AF HYB is used to connect a local transmitter and receiver to a two-wire telephone communication line. The module provides isolation between the transmitter and receiver through its hybrid network.

Accessible from the front of the board are six jacks to allow patching into the module. These provide for testing the line or equipment at the transmitter and receiver ports. Jacks are also available for testing the equipment and for monitoring the two-wire communication line.

An internal balancing network is utilized to optimize hybrid loss. If the internal network does not provide an adequate trans-hybrid loss, an external balancing network may be connected.

An optional talk-battery feed coil is available. It is capable of handling 100 mAdc. Optional balanced H pads can be ordered to reduce signal levels to the transmitter and from the receiver. 6-dB Receive, and 12-dB-Send pads are standard. Other levels of attenuation are available.

SPECIFICATIONS

Frequency Response

Between +0.1 and –0.5 dB, relative to the response at 1 kHz, over the range from 0.3 to 6 kHz.

Transhybrid Loss

40 dB minimum from 0.3 to 6 kHz without battery-feed coil. 33 dB minimum from 0.3 to 6 kHz with battery-feed coil. These can be improved by using an external balancing network.

Insertion Loss

Less than 4.0 dB.

Return Loss

ERL: 27 dB minimum, without battery-feed coil. 25 dB minimum with battery-feed coil at 100 mA.

SRL_{LO}: 20 dB minimum, without battery-feed coil. 18 dB minimum with battery-feed coil at 100 mA.

Longitudinal Balance

60 dB minimum from 0.3 to 6 kHz.

Third-Harmonic Distortion

51 dB minimum, with a 300-Hz signal at +10 dBm.

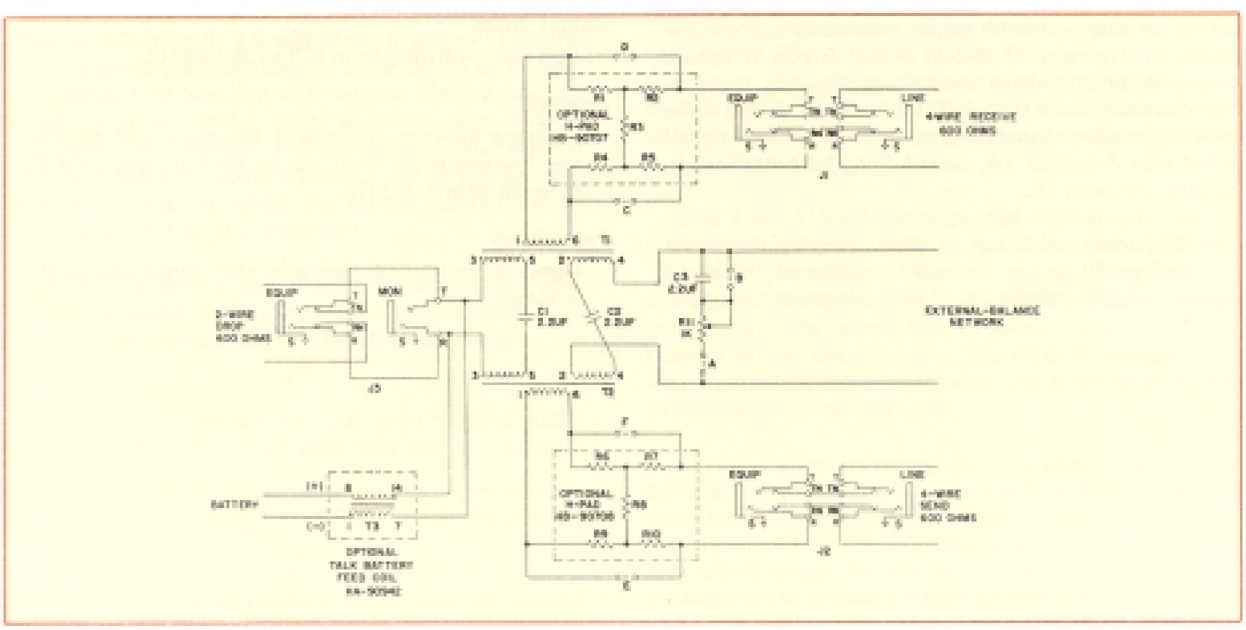


Figure 5.27. Schematic of circuit, Model 65 AF HYB.

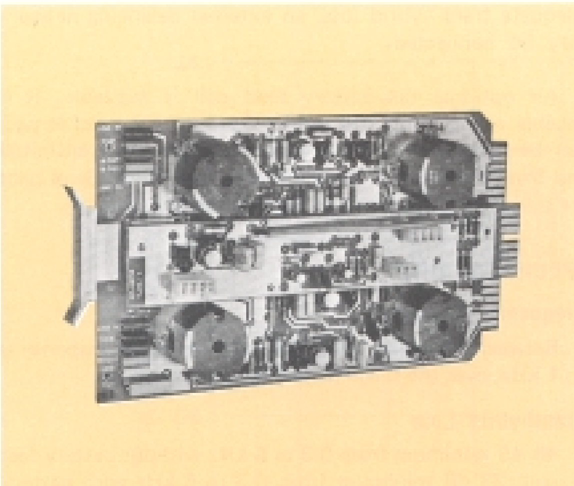


Figure 5.28. Model 68 WB DUAL AMP Dual Wideband Amplifier.

68 WB DUAL AMP WIDEBAND AMPLIFIERS

The Model 68 WB DUAL AMP Wideband Dual Amplifier consists of two two-section amplifiers and an optional multiplexer which functions as a summing amplifier. Each of the signal amplifiers consists of an input-amplifier section and an output-amplifier section. Many different applications and operating levels are achievable by simple placement of the many jumpers provided and by arrangement of external connections. The Multiplex Adapter, Option HB-46125, operates over a frequency range of 300 to 60,000 Hz, and it is built on a plug-on circuit board that fits over the main circuit board.

This module can be used as a general-purpose amplifier and combiner for audio, or for amplifying and combining express-order-wire voice-carrier signals operating in the range between 4 and 60 kHz. The optional multiplex adapter can be used in conjunction with Amplifiers 1A and 1B for single-sideband carrier transmission. The transmission may consist of Group A and Group B signals, along with synchronizing carrier and signaling tones. A voice channel from 300 to 4000 Hz is also combined at the multiplex adapter's input, and all of the combined signals are fed into Amplifier 1A, passed along to Amplifier 1B, and then out from 1B.

In the opposite direction, or when receiving, the output of Amplifier 2B drives the audio portion of the signal through output transformer T401. However, if single-sideband signals are being received, then Jumper J403 is removed and the composite signals are sent to the buffer amplifier of the multiplex adapter. The buffer amplifier's output is fed to a load-matching-resistor network, which matches the input circuit of subsequent demodulator filters with an input impedance of 75 ohms.

Provision has been made at Input Amplifier 2A for baseband coupling. Provision for the installation of a T pad is at the output of each output amplifier, with the choice of attenuation determined in accordance with the requirements of the application.

SPECIFICATIONS

Input and Output Impedance

Balanced or unbalanced, 600 ohms, with or without transformers. Return loss is greater than 20 dB in the frequency range between 300 and 30,000 Hz.

Frequency Response

With transformers: within 0.5 dB of 1000-Hz reference from 300 to 30,000 Hz.

Gain

Continuously adjustable from −30 to +35 dB, using jumpers to set the range of a multi-turn potentiometer.

Total Harmonic Distortion

Less than 0.2% from 4 to 30 kHz. Less than 1.5% from 200 to 4000 Hz, all at maximum output level.

Maximum Output Level

+10 dBm.

Output Noise

Less than −65 dBm from 200 to 30,000 Hz.

Crosstalk

Coupling between channels is less than −70 dB from the low-frequency limit up to 10 kHz, and −60 dB between 10 and 30 kHz.

Front Panel

Optional front panel available with input and output telephone jacks.

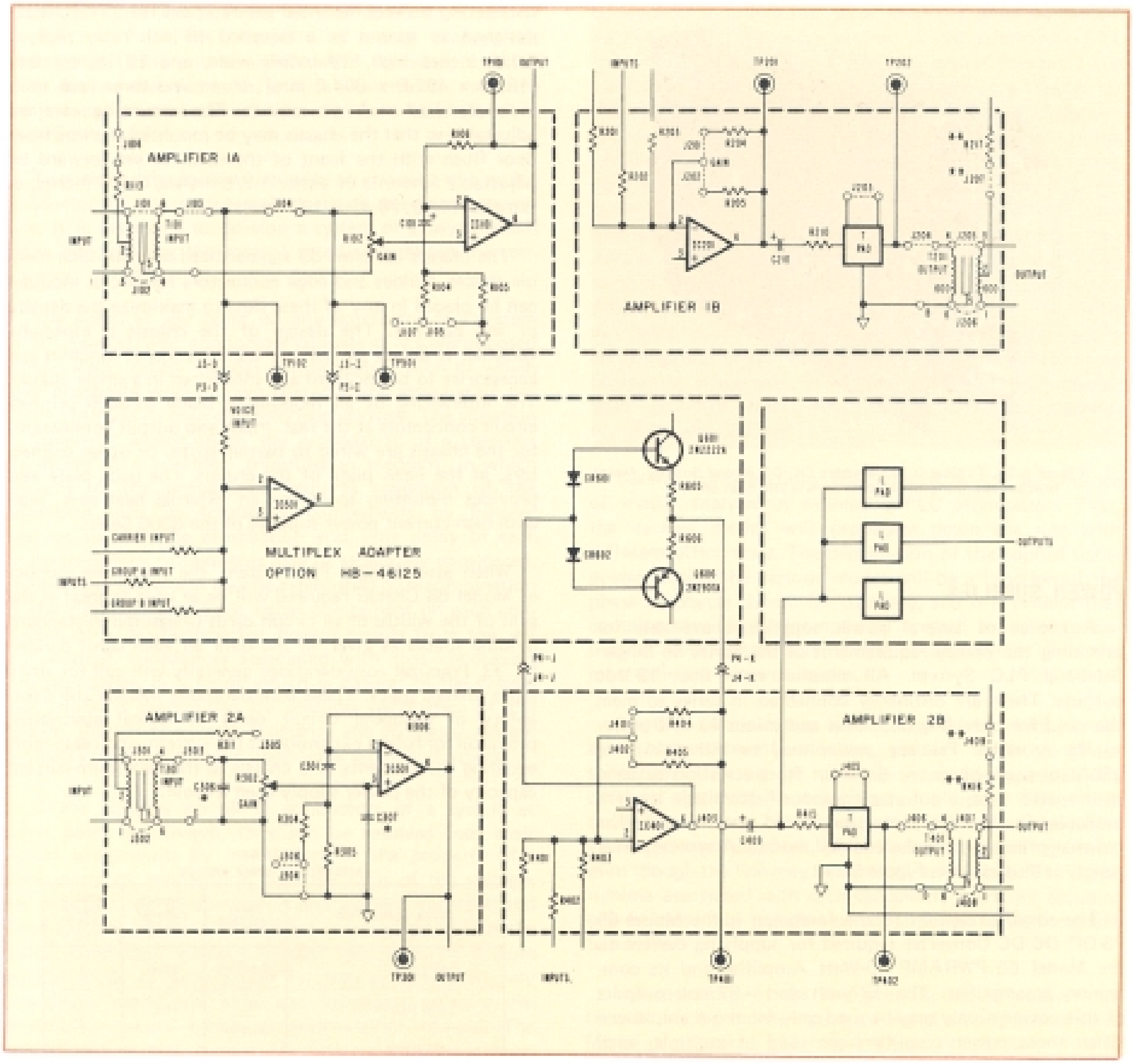


Figure 5.29. Functional schematic of dual wideband amplifier with optional multiplex adapter.

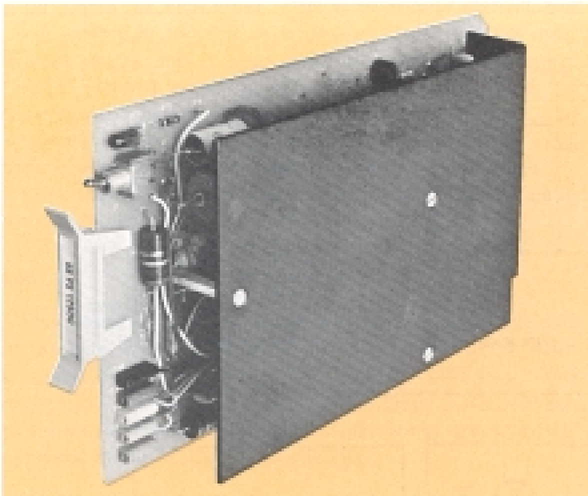


Figure 5.30. Typical 0.375-ampere DC-DC Power Supply. Metal bracket spanning the circuit card is the heat sink.

POWER SUPPLIES

A choice of several power supplies is available for providing the energy requirements of the Series 65 Single-Sideband PLC System. All supplies have dual 12-Vdc outputs. These are ordinarily connected in series to meet the need for providing both plus and minus 12-volt power to the modules. Primary power may be either 115- or 230-Vac, single-phase, or dc input from a station battery. In the latter case, input at a number of dc-voltage levels is available. Overvoltage protection, based upon the familiar crowbar principle, is included in all models. A typical power supply is illustrated in Figure 5.27.

The single exception to the foregoing is the Model 65 PS-DC DC-DC Converter required for supplying current to the Model 65 PWRAMP 40-Watt Amplifier and its companion preamplifier. The +37-volt and −37 volt outputs of this power supply may be used only for those amplifiers. When these power amplifiers are used in multiple, each must have its own power supply.

Table 5.1 is a listing of power supplies available for use in the Series 65 System. More than one may be required, either to supply the total current needed, or to function as a standby unit to improve reliability of the system.

MECHANICAL CONSIDERATIONS

Because the RFL Series 65 is not a standardized system offering only one modulation plan and a fixed number of channels, the designer is free to establish the PLC system desired, providing for growth as expected, and to select the type and quantity of individual circuit modules needed to create the system. This freedom extends also to the mechanical aspects of the design.

The basic enclosure for the Series 65 Single Sideband Carrier System is the Model 68 Chassis. Into the basic unit, individual circuit cards plug in, bookshelf-style, into connecting sockets mounted inside at the rear. The chassis, designed to mount in a standard 19-inch relay rack, is 5-1/4 inches high, 19 inches wide, and 12 inches deep (133.4 x 482.6 x 304.8 mm). It requires three rack units in a standard rack or cabinet. The mounting ears are adjustable so that the chassis may be mounted with its front door flush with the front of the rack, or set forward by adjustable amounts of either 1-7/8 inches, 3-1/8 inches, or 5 inches (46.9, 79.4, or 127 mm).

The chassis contains 33 standardized one-half-inch module spaces. Guides and edge connectors for circuit modules can be placed in any of these slots to maximize the density of the package. The design of the chassis is especially versatile. It allows almost any combination of modules and accessories to be included and interwired in a single chassis. The interwiring of all modules is made through printed-circuit connectors at the rear. Input and output connections for the chassis are wired to barrier strips, or other connectors, at the back plate of the chassis. The back plate also provides mounting space for an external heat sink, used with high-current power supplies of the 6000 Series.

When assembling a PLC system, the minimum number of Model 68 Chassis required will be at least as great as the sum of the widths of all circuit cards (measured in standard module spaces as given in the data on each card) divided by 33. Practical considerations generally will call for using more chassis space. These considerations include the desirability of grouping circuit cards in logical assemblies, provision for future expansion of the system, and reasonably equating the contents of a chassis to the maximum-current capacity of the power supply used therein.

TABLE 5.1
RECOMMENDED POWER SUPPLIES

MODEL DESIGNATION	INPUT VOLTAGE	OUTPUT CURRENT (1) (3)	MODULE SPACES REQ'D (2)
68PS 12DC-1	10.5-14 Vdc	0.375 A	3
68PS 24DC-1	21-28 Vdc	0.375 A	3
68HPS 24DC-1		1.0 A	
68PS 48DC-1	42-56 Vdc	0.375 A	3
68HPS 48DC-1		1.0 A	
68PS 125DC-1	104-140 Vdc	0.375 A	3
68HPS 125DC-1		1.0 A	
68PS AC-1	115/230 Vac ±10%	0.25 A	3
68A HPS AC-1	48-68 Hz	1.0A	5
65PS 48DC	42-56 Vdc	Dual 12-Vdc 2.0-amp, and Dual 37-Vdc 1.0-amp. (4)	Relay-rack mounted
65PS 125DC	108-138 Vdc		

(1) This is output current for each of the two independent 12-Vdc output circuits. Because these are usually series connected to provide both plus and minus outputs, this will ordinarily be the total current capacity of the power supply.

(2) This is the number of standard module spaces occupied by the power supply when plugged into a Model 68 Chassis which contains a total of 33 one-half-inch module spaces.

(3) One-ampere units require space at rear of Model 68 Chassis for mounting a heat sink, for which provision is made on the chassis.

(4) Dual outputs are series connected to provide both plus and minus voltages at the current capacity indicated.

A chassis with circuit modules installed is shown as Figure 5.31. Mounting and outline dimensions are shown in Figure 5.32.

The quantity of Model 68 Chassis used to assemble a PLC terminal may be mounted together in any relay rack or relay-rack cabinet that suits the requirements of the installation. A typical assembly is shown in Figure 5.33. Outlines of two types of cabinets available from RFL are shown in Figure 5.34. The smaller of these is a NEMA Type 4 enclosure, useful in areas where serious problems of corrosion or weather exist. Other styles and sizes of enclosures also are available.

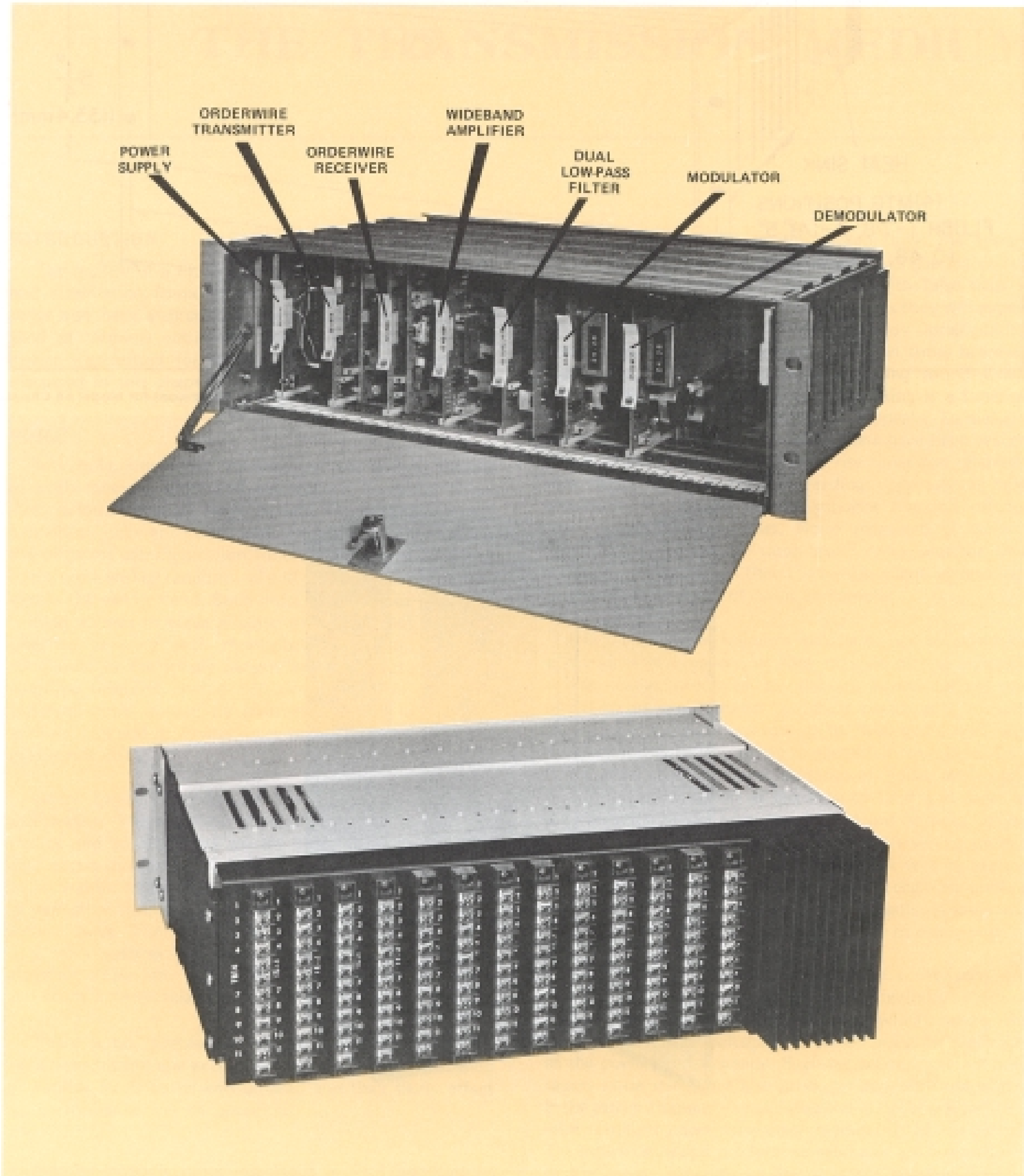


Figure 5.31. Typical Model 68 Chassis with circuit cards installed.

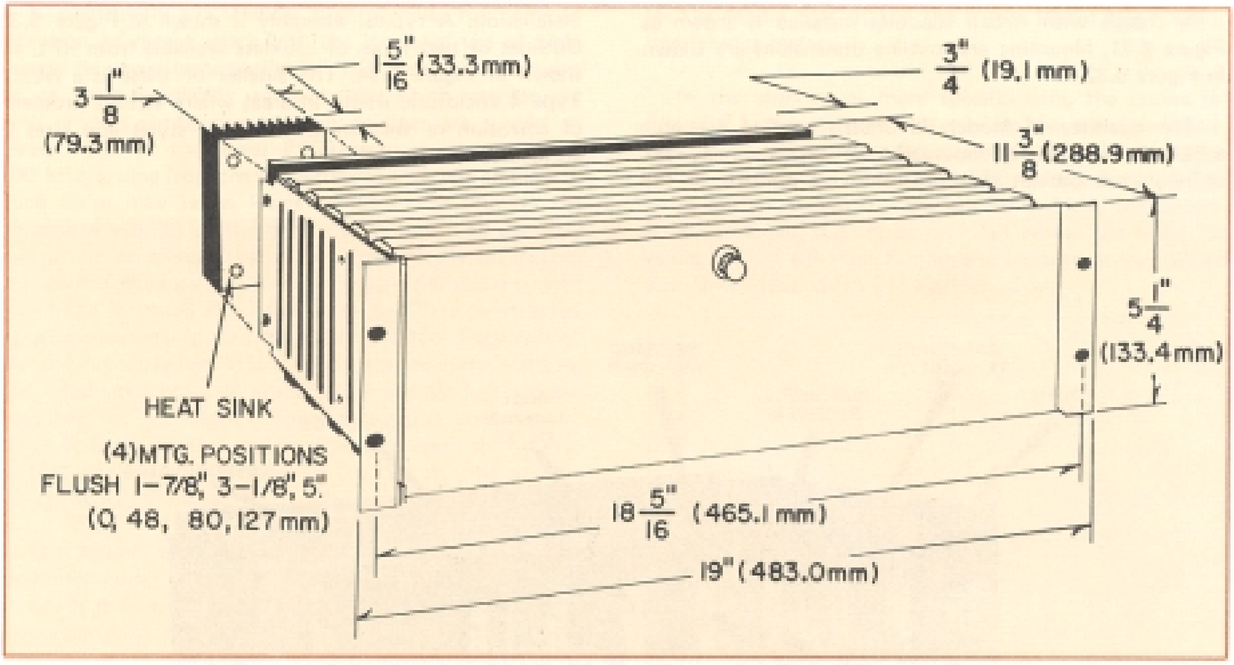


Figure 5.32. Mounting and outline dimensions for Model 68 Chassis.

Figure 5.33. Typical assembly of equipment for a PLC terminal.

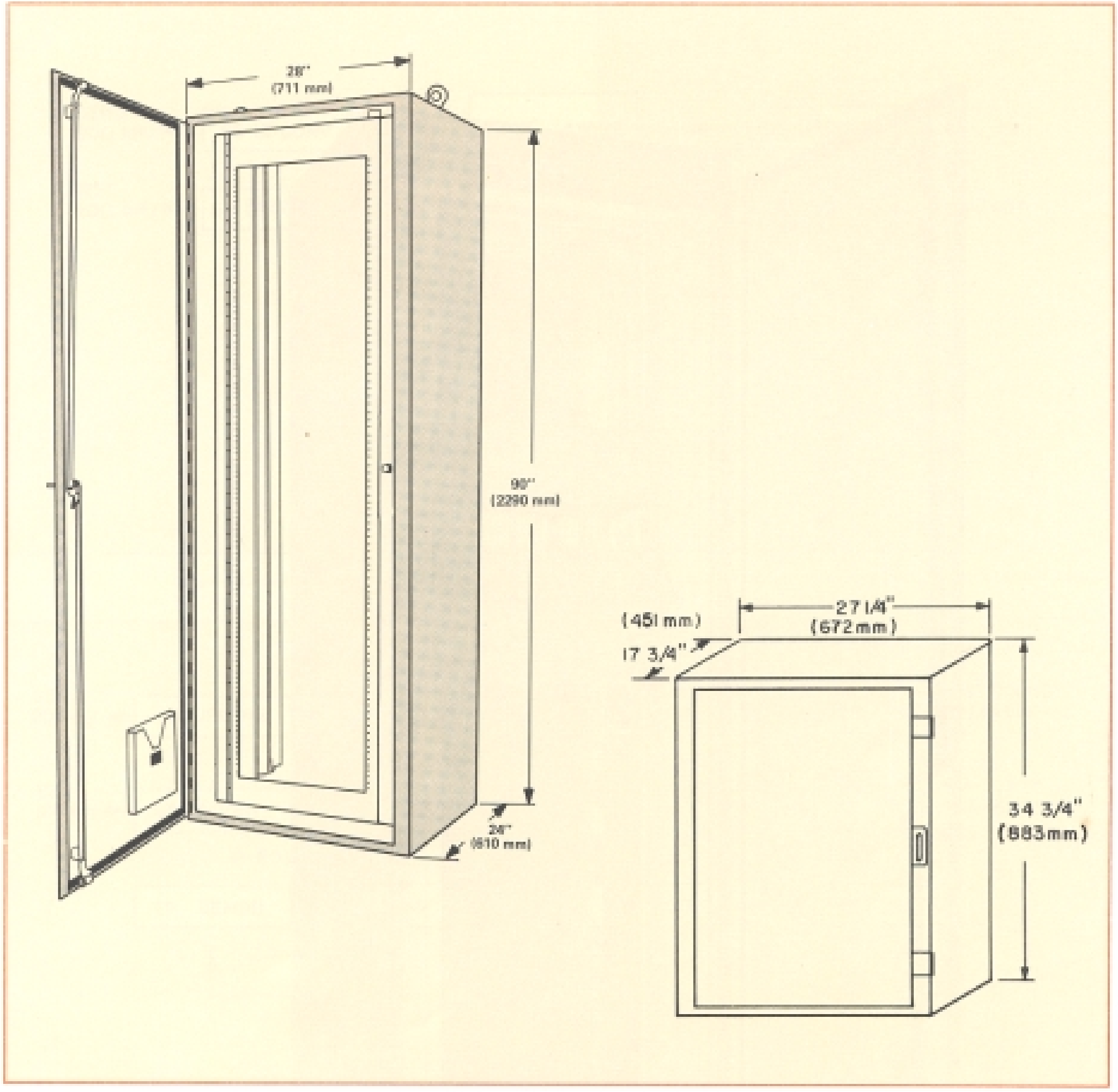
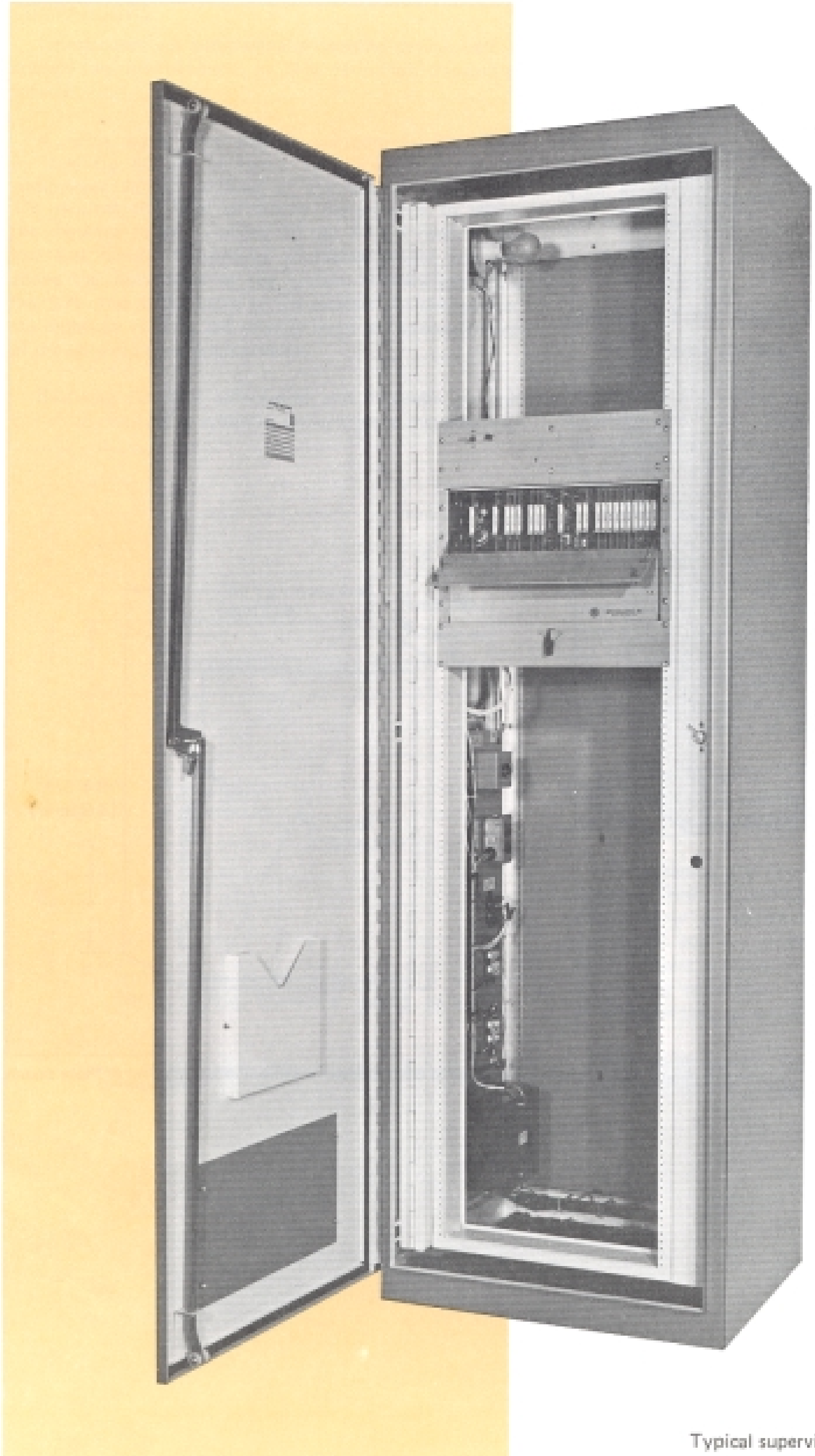


Figure 5.34. Outline Dimensions of Basic Enclosure without Mounting Feet or Floor Stands.



Typical supervisory-control terminal operated over a PLC system.

Chapter 6

COUPLING EQUIPMENT

INTRODUCTION

Line traps, inductors, coupling and bypass capacitors, and capacitive voltage transformers are used to terminate and to couple to and from the powerline at carrier-current frequencies. These devices can be purchased from RFL as part of a PLC system, and details are given in this section. The products of Trench Electric Limited exhibit a wide variety in such equipment, and they have been chosen for description as typical of what is available.

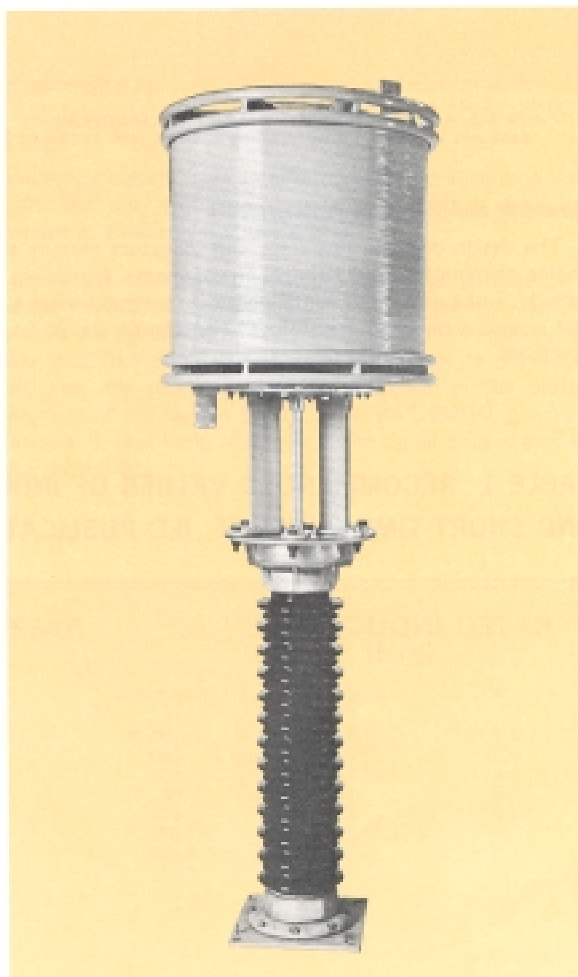


Figure 6.1. Line trap mounted on a coupling capacitor with a custom-made pedestal.

TRAPS and INDUCTORS

Line traps and inductors are available with rated continuous current and short-time currents corresponding to the requirements of the International Electrotechnical Commission (IEC Publication 353), and of the American National Standards Institute (ANSI C93.3). Special designs are available to meet other standards and special requirements. Typical line traps are shown as Figures 6.1 and 6.2.

Detailed information is available from RFL.

Application of Traps and Inductors

The basic function of a carrier-current line trap is to insert a high impedance in the transmission line at a selected frequency or bands of frequencies. This impedance is used to prevent grounding and loss of the PLC signal during fault conditions and to reduce attenuation of the signal by the station bus behind the line trap. Design considerations are discussed in Chapter 2.

Because the trap is inserted in series with the transmission-line conductor, it must carry the power-frequency current. Moreover, it must be designed to withstand the mechanical forces created by short-circuit currents associated with the transmission system.

Overload Capabilities

Line traps and inductors must be capable of withstanding overload conditions without damage. The percentage overload permitted will be a function of ambient temperature and type and rating of the coil. Figure 6.3 shows a typical overload curve for three ambient temperatures. Specific overload ratings are available on request. See also Table 1.

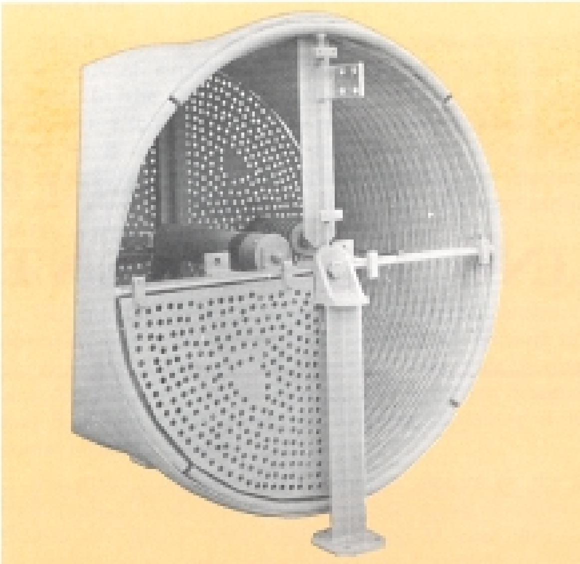


Figure 6.2. A typical horizontal-pedestal line trap with tuning pack and lightning arrestor. Rating is 0.265 mH, 1200 amp.

Mounting and Terminal Arrangements

The design of these line traps and inductors permits a choice of mounting and terminal arrangements. Mountings include suspension, horizontal-pedestal, vertical-pedestal, and coupling-capacitor mountings. The choice should be specified.

A considerable saving often can be achieved by using a single vertical pedestal to allow the trap to be supported with just one insulator stack. This method also offers the advantage of using a simpler mounting structure. Line traps with a vertical-pedestal mounting often may be mounted on top of a coupling capacitor, without further support, depending on the ratings of the trap and capacitor. Figures 6.1 and 6.4 are illustrative.

For horizontal-pedestal mounting, two methods are available. Larger inductors are mounted with the axis of the coil in a vertical position, to allow proper cooling. To conform to requirements of the installation, an aluminum channel is supplied to provide for two-insulator mounting schemes. Figure 6.5 shows an example. Smaller traps can be mounted in a horizontal position as shown in Figure 6.6. Terminals can be positioned on any spider member of the coil. This allows maximum flexibility for bus and cable connections.

COUPLING CAPACITORS and CVT's

Coupling capacitors provide a means for coupling to and from the high-voltage transmission line. These capacitors can also be supplied with voltage-transformer circuits to provide operating power for terminal equipment at the site. When so equipped, the device is known as a capacitive voltage transformer (CVT). A capacitive voltage divider is formed and the tap voltage is applied to a reactor and transformer circuit to give an output at the required voltage and phase. Typically, CVT's will supply maximum burdens in the range from 200 to 800 volt amperes.

TABLE 1 RECOMMENDED VALUES OF INDUCTANCE, CONTINUOUS CURRENT, AND SHORT-TIME CURRENT IEC PUBLICATION 353 AND ANSI C93.3

RATED INDUCTANCE (mH)		RATED CURRENT IEC			RATED CURRENT ANSI	
IEC	ANSI	Continuous	Short Time Series 1	Short Time Series 2	Continuous	Short Time
0.2 mH	0.265 mH	100 A	2.5 kA	5 kA	400 A	15 kA
0.25	0.53	200	5	10	800	20
0.4	1.06	400	10	16	1200	36
0.5	1.59	630	16	20	1600	44
1.0	2.12	800	20	25	2000	63
2.0	2.65	1,000	25	31.5	3000	63
		1,250	31.5	40	4000	80
		1,600	40	50		
		2,000	40	50		
		4,000	50	63		

Rated Inductance is measured at 100 kHz per ANSI C93.3
at 1.0 kHz per IEC Publication 353

Rated Short-Time Current: r.m.s. for one second. Mechanical rating is r.m.s. x 2.55 for first current peak.

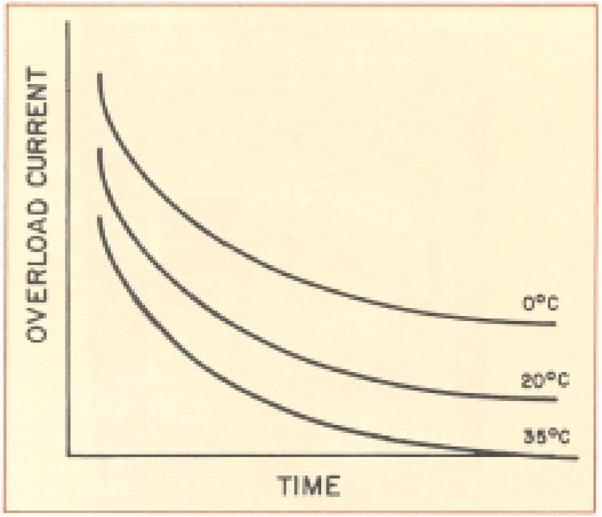


Figure 6.3. Typical overload characteristics for line traps and inductors.

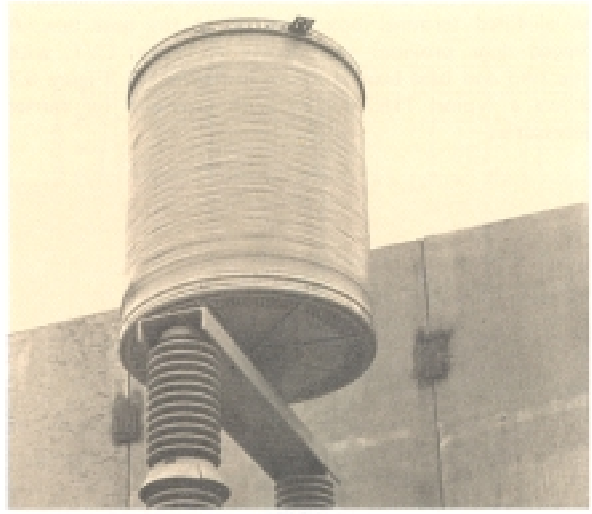


Figure 6.5. Line trap, rated at 2000 amperes, 0.35 mH, 230 kV, using a horizontal-channel mounting. This keeps the axis vertical for optimum cooling.

Types

Types TEV, TEM, and TEHM are capacitor voltage transformers used for powerline-carrier coupling and/or relaying and metering applications, while Types TEC and TEHC are coupling capacitors. Typical electrical performance and dimensional data are shown in Tables 2 and 3. A typical CVT is illustrated in Figure 6.7.

Design details are available from RFL.

Construction

Coupling capacitors consist of a series of capacitor elements housed in a high-strength, oil-filled porcelain shell. They follow a flat, foil-wound, low-loss design using a high grade of thin Kraft paper. Capacitor elements are dehydrated under vacuum and impregnated with mineral oil also purified and degassed under vacuum. Each capacitor unit is fitted with a stainless-steel expansion chamber, and the chamber uses a sharpened steel puncture pin to release oil in case of pressure build-up. Specific voltage ratings are obtained by stacking one or more individual units.

Figure 6.4. Line trap mounted on capacitor voltage transformer.

Figure 6.6. An 800-ampere, 0.265-mH line trap being installed in the horizontal position.

The lowest capacitor section is fitted on a welded-steel base box. For CVT's the box will house the intermediate transformer, compensating reactor, and associated circuits. Safety and reliability are enhanced by enclosing the intermediate-voltage circuit components in insulating oil and protecting them from environmental deterioration. Although operable without oil, the insulation safety factor is increased greatly by using this system. The secondary terminals, protective device and carrier accessories are contained in an air-filled terminal box attached to the base box. A hinged door provides easy access. A typical CVT, with line trap and base box, is shown in Figure 6.4. Figure 6.7 shows a typical 115-kV CVT with provision for carrier accessories.

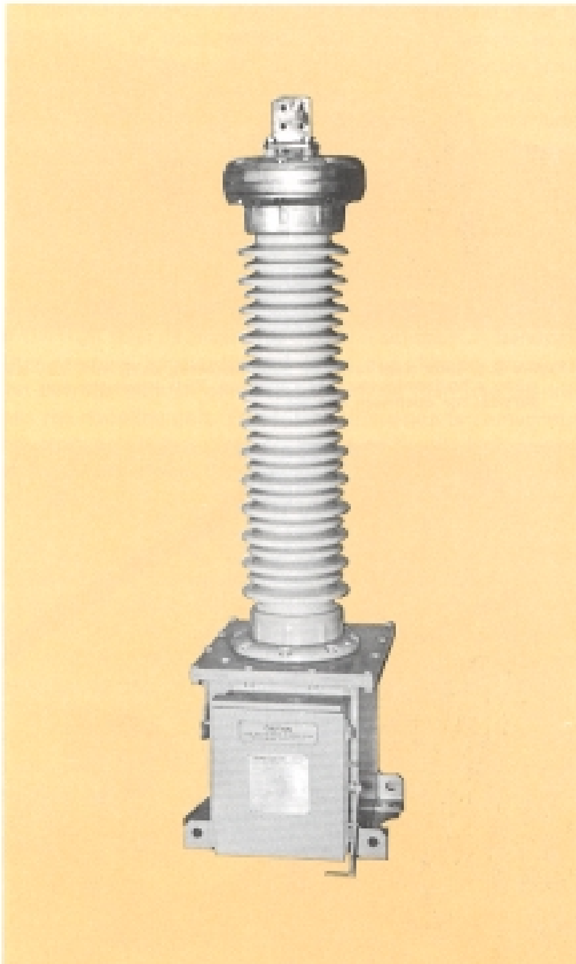


Figure 6.7. Type TEV 115 Capacitor Voltage Transformer with high-voltage terminal and carrier accessories.

Carrier Accessories

Carrier accessories include a ground switch, drain coil, and protective gap, plus a fitting for either high-impedance cable or coaxial cable. Carrier accessories can also be supplied separately for modification of units already in service. Line-tuning units, or filters, can be supplied mounted in a separate weatherproof box. For Types TEC and TEHC, line-tuning units may be mounted in the base box of the coupling capacitor. Detailed information is available from RFL. A typical terminal box is shown as Figure 6.8.

Mounting

Coupling capacitors and CVT's can be pedestal- or suspension-mounted. Mounting dimensions can be varied,

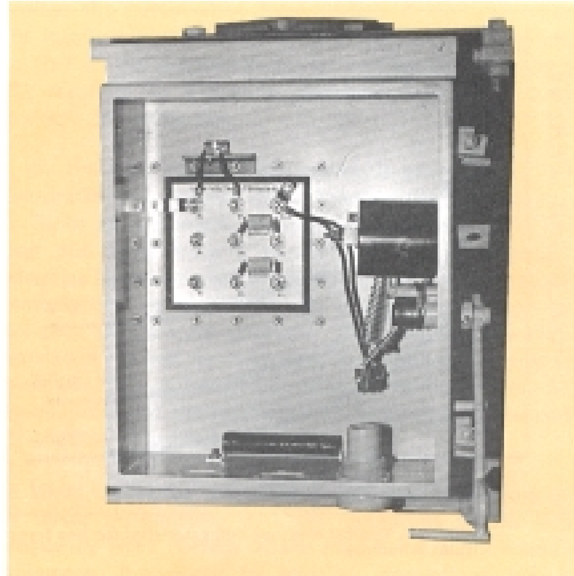


Figure 6.8. Low-voltage terminal box, showing carrier accessories and carrier-entrance bushing.

in many cases, to allow direct substitution with other manufacturer's units, thereby giving the advantage of flexibility. A standard high-voltage NEMA four-hole aluminum blade terminal or a single, aluminum, stud-type terminal is supplied. This terminal can be removed to accommodate a standard line-trap mounting using a bolt circle five inches in diameter.

The use of high-strength, wet-process porcelain for capacitor shells permits vertical mounting of line traps. Many installations can use this type of mounting, with a considerable reduction in cost. Consideration must be given to wind loading, seismic requirements, voltage ratings, and size of line traps. Values of permissible horizontal pull, and other design details are available from RFL.

LINE-TUNING UNITS

Line-tuning units, based on the principles outlined in Chapter 2, are available in the following forms:

(a) Single-frequency tuners, which pass just one frequency. Generally their bandwidth is useful up to 12 percent of the resonant frequency.

(b) Two-frequency tuners, in which two frequencies, separated by at least 25 percent of the higher frequency are passed.

(c) Wideband tuners, in which a group of frequencies, whose bandwidth is determined largely by the value of the coupling capacitor, are passed.

(d) Resonant bypass tuners, which provide a through path for PLC signals around discontinuities such as circuit breakers, power transformers, or lines of different voltage.

Details on line tuners are available from RFL.

TABLE 2 ELECTRICAL PERFORMANCE

TYPE	ACCURACY and BURDEN	TRANSIENT RESPONSE*	FERRORESONANCE SUPPRESSION*	No. OF SECONDARIES
TEV	ANSI-0.6WXYZ on each winding. Total simultaneous burden — 200VA (0.6 class) IEC-150VA Cl 0.5 on each winding. Total simultaneous burden — 150VA Cl 0.5 Maximum Output — 600VA (thermal limit)	Less than 10% (of crest) residual voltage in 1 cycle	Less than 10% (of crest) in 200 milliseconds at 120% of rated voltage	Standard — 2 (optional 25VA third winding).
TEM	ANSI-0.3WXYZ on each winding. Total simultaneous burden — 200VA (0.3 class) IEC-200VA Cl 0.5 on each winding. Total simultaneous burden — 200VA Cl 0.5 Maximum Output — 600VA (thermal limit)	Less than 10% (of crest) residual voltage in 1 cycle	Less than 10% (of crest) in 200 milliseconds at 120% of rated voltage	Standard — 2
TEHM	ANSI-0.3WXYZZ on each winding, 1.2Y auxilary winding. Total simultaneous burden — 400VA (0.3 class) IEC-400VA Cl 0.5 on each winding. Total simultaneous burden — 400VA Cl 0.5 Maximum output — 800VA (thermal limit)	Less than 5% (of crest) residual voltage in 10 milliseconds.	Less than 10% (of crest) in 200 milliseconds at 120% of rated voltage	Standard — 2 (optional 75VA third winding)

*Test results can be provided upon request.

TABLE 3 ELECTRICAL PARAMETERS AND DIMENSIONS

TYPE	HIGHEST SYSTEM VOLTAGE LINE TO LINE kV	BIL 1.2/50us (kV)	POWER FREQUENCY TEST LEVEL (kV)	CAPACITANCE (pF)[3]	HEIGHT A INCHES/MM
TEV, TEM, TEC					
69	72.5	350	165	10,000 (11500)	54.5/1380
115	123	550	265	6,000 (6800)	68.5/1740
138	145	650	320	5,000 (5700)	77/1950
161	170	750	370	4,300 (4900)	101.5/2568
230	245	1050	525	3,000 (3400)	115.5/2940
230H	245	1050	525	3,000 (3400)	133/3370
300	300	1300	640	2,300 (2600)	149/3780
345	362	1550	785	2,000 (2300)	163/4140
400	420	1550/1800	785	1,700 (1900)	[1]188.5/4790
500	550	1800	900	1,500 (1750)	209.7/5305
TEHM, TEHC					
115	123	550	265	20,000	72.3/1836
138	147	650	320	17,500	80.9/2057
161	170	750	370	14,000	88/2235
230	245	1050	525	10,000	122.8/3119
345	362	1550	785	6,700	173.3/4401
400	420	1550/1800	785	5,800	[2]199.1/5057
500	550	1800	900	5,000	224.5/5702
765	765	2200	1200	4,000	325/8255

[1]1800kV BIL unit height — 210/5330
[2]1800kV BIL unit height — 224/5689
[3]Capacitance of TEM only in brackets

For further details of technical data refer to Bulletin T300-05-15.

Data not to be used for construction purposes.

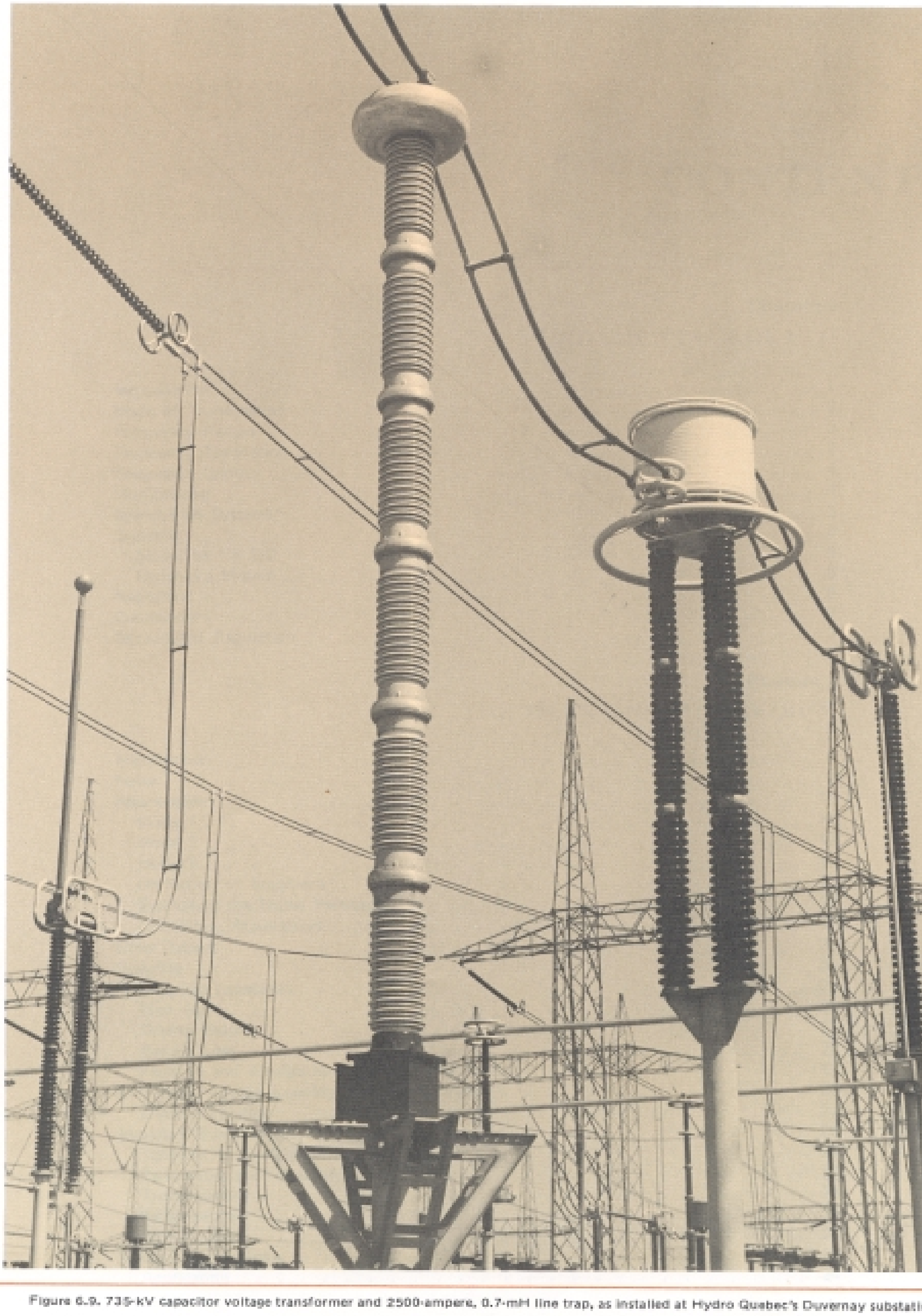


Figure 6.9. 735-kV capacitor voltage transformer and 2500-ampere, 0.7-mH line trap, as installed at Hydro Quebec's Duvernay substation.

RFL

EST. 1922

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