AN INVESTIGATION INTO POTENTIAL AM BROADCAST RECEIVER DESIGN IMPROVEMENTS AND THEIR POSSIBLE IMPACT ON SPECTRUM UTILIZATION

MILLER COMMUNICATIONS

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FINAL REPORT

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1.0 INTRODUCTION

1.1 Purpose Of The Study

In ITU (International Telecommunication Union) Region 2 (North and South America, Greenland), the common AM broadcast band is currently allocated the frequency spectrum extending from 535 kHz to 1605 kHz. At the present time, there are more than 600 broadcast stations using the 107 channels, each 10 kHz wide, which comprise the AM band. This high degree of congestion has been of growing concern - especially in North America - over the years because of the great difficulty involved in assigning AM channels to new broadcast stations so that they might still be able to provide an interference-free service to their associated listening populations.

Accordingly, a proposal for the extension of the upper frequency limit of the AM band from the current 1,605 kHz to 1,705 kHz was submitted by Canada to the World Administrative Radio Conference (WARC) which was held in Geneva in September 1979.

In the WARC decision (Recommendation YC), broadcasting was granted exclusive use of the band up to 1,625 kHz and would be the primary user of the band from 1,625 to 1,705 kHz. Actual use of the band between 1,625 kHz and 1,665 kHz would not commence, however, until after July 1987; for the band 1,665 kHz to 1,705 kHz, usage would not commence before July 1990. The necessary frequency planning for the use of this extended band* will be the prime focus of the next Regional Administrative MF Broadcasting Conference (Region 2) to be held in 1981. As part of Recommendation YC, administrations of Region 2 were encouraged to "promote the

*In the Final Acts of WARC 79 (Geneva), AM broadcasting was also allowed usage of the 525 to 535 kHz band on a shared basis with Aeronautical Radionavigation.

development and availability of receivers suitable for the broadcast band extended to 1,705 kHz".

It is with this exhortation in mind and the possibility of AM receiver design and performance modifications engendered by the impending band extension as well as other factors (e.g., reduction of the AM channel spacing from 10 kHz to 9 kHz) that this study has been undertaken. The objectives are to determine what, if any, impact the extension of the band will have on present AM receiver designs as well to examine other facets of receiver design which might possibly promote more efficient utilization of the AM frequency spectrum.

1.2 Methodology

In preparing this report, supporting material has been culled from published texts, technical literature, manufacturers' product literature and in-house design knowledge.

As foreground material to the discussions on the AM broadcast receiver that follow, Section 2 commences with a discussion of the signal format used in AM broadcasting, the spectral occupancy of the AM band and the location of its image band for a typical receiver intermediate frequency (IF) of 455 kHz. The propagation characteristics of AM signals is next discussed and points out the difference between daytime and nighttime propagation. Because of the mutual interference which can be experienced by stations either operating on the same channel (co-channel interference) or on adjacent channels (adjacent channel interference) both during the day and at night, fairly strict rules have been formulated to govern the maximum acceptable level of interference. For the North American region, these rules were embodied under a regional agreement known as the North American Regional Broadcasting Agreement (NARBA) in 1950. Section 2 concludes with a brief discussion of the NARBA classifications for broadcast stations, broadcast channels and the various co-channel and adjacent channel interference protection afforded to the various classes of stations.

Section 3 constitutes the major part of this study in dealing with the design aspects of the AM broadcast receiver. The Section commences with a synopsis of the CCIR specifications relating to the minimum recommended performance which AM receivers are expected to achieve consistent with the CCIR's objective of promoting large-scale production of low-cost receivers which might be made more available to areas of the world with low receiver-population densities. A brief description of the typical AM broadcast receiver is then followed by a detailed discussion of the design aspects of each facet of the AM receiver starting from the antenna and RF stage to the baseband detection stage wherein the original transmitted information is recovered. The only feature omitted from these discussions is the audio driver/power amplifier stage since it has no direct bearing on the manner in which the AM spectrum might be more efficiently utilized. In each of the relevant sub-sections and in the concluding portion of Section 3, the specific design issues raised in the Statement of Work are discussed.

Section 4 (Conclusions) highlights the specific receiver design issues relating to spectrum utilization efficiency and deliberates on the relative acceptability of cost increases which might result from design changes for the three classes of receivers discussed, namely, the low-cost

($\sim$\$5) portable pocket receiver, the medium cost ($\sim$\$70) portable A.C./D.C. receiver and the premium cost ($\sim$\$140) console-type receiver.

Finally, Appendix A surveys some of the common passive IF filters available today in order to evaluate their suitability for application in the AM receiver's IF stage on the basis of performance, size and cost. The filter types discussed include the popular LC single/double-tuned IF transformer, the crystal filter, the mechanical filter, the Surface Acoustic Wave (SAW) filter and the ceramic filter.

2.0 AM BROADCASTING

2.1 Signal Format

In commercial AM broadcasting, the form of modulation employed is amplitude modulation of an unsuppressed carrier of constant amplitude yielding two symmetrical sidebands, i.e., double sideband large carrier transmission (DSB-LC) or simply amplitude modulation (AM), as it is more commonly known. Since the carrier of an AM signal conveys no baseband information, a substantial amount of power is wasted in its transmission but it is the price which must be paid if AM broadcast receivers, of which there are millions, are to remain inexpensive.

The amplitude modulated (AM) carrier signal may be mathematically represented as:

$$\phi_{AM}(t) = [A_c + f(t)] \cos \omega_c t \qquad (2.1)$$

where A_c = amplitude of the unmodulated or pure carrier signal

$f(t)$ = baseband modulating signal

$f_c = \frac{\omega_c}{2\pi}$, carrier frequency.

Written in this form, $\phi_{AM}(t)$ may be regarded as a carrier signal $\cos \omega_c t$ whose amplitude is $[A_c + f(t)]$. Figure 2.1 illustrates the time and frequency domain representations of $\phi_{AM}(t)$ and its various component terms.

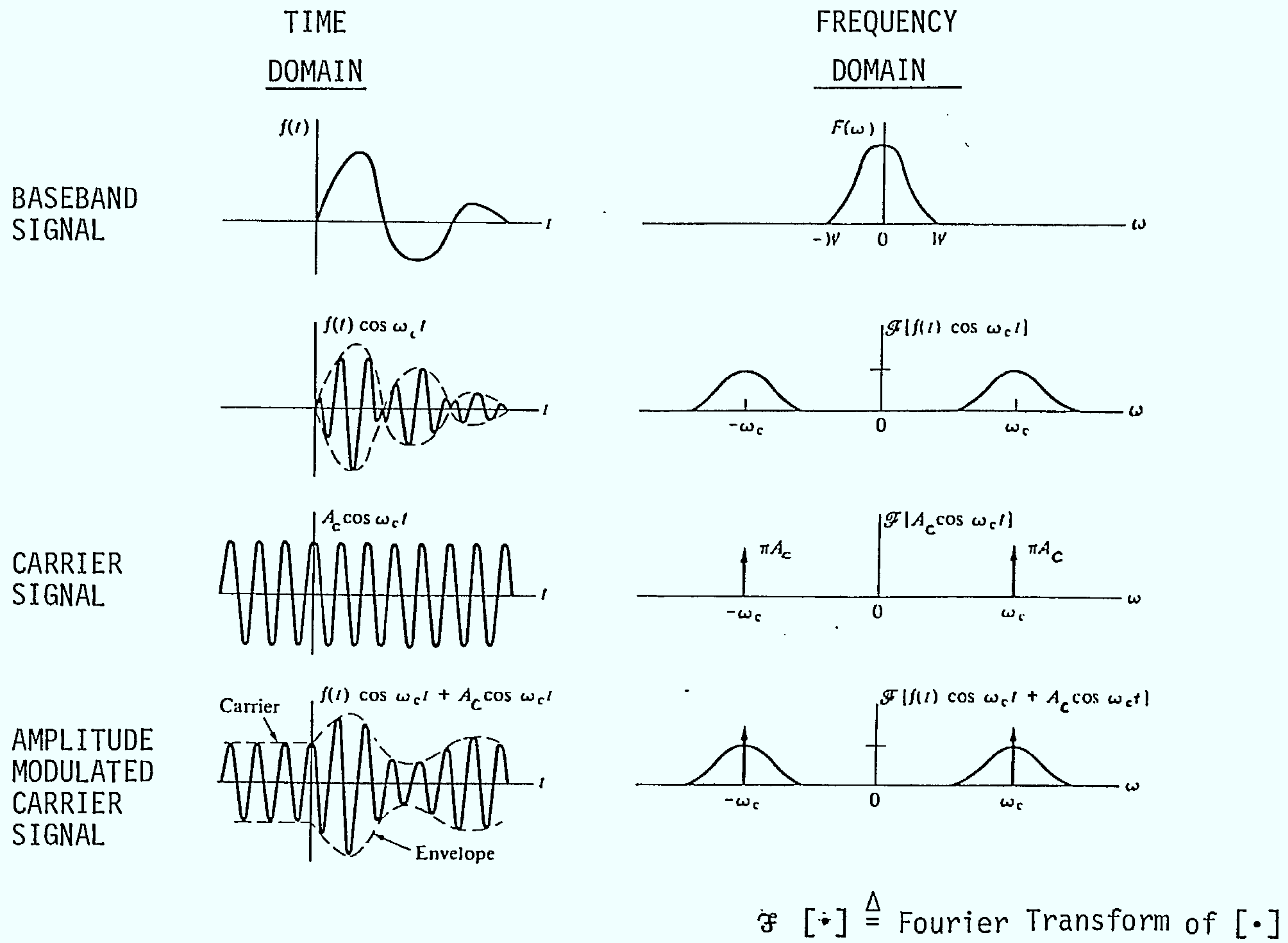


FIGURE 2.1 TIME AND FREQUENCY DOMAIN REPRESENTATIONS OF AN AM BROADCAST SIGNAL

In Figure 2.2(a), we note that provided the pure-carrier amplitude A_c is large enough, the envelope of the modulated carrier will be proportional to f(t) and hence demodulation of an AM signal reduces to the detection of the carrier envelope, with no dependence on the frequency or phase of the carrier. If, however, f(t) is allowed to exceed A_c, the carrier envelope will no longer be proportional to f(t) (Figure 2.2(b)) and severe baseband distortion will result if an envelope detector (see Section 3.6.1) is being used. A synchronous detector (Section 3.6.3), however, will still correctly demodulate such signals.

Thus for simple envelope detection, which is universally employed in AM broadcast receivers, it must be ensured that the amplitude, A_c, of the unmodulated (pure) carrier signal is always greater than or equal to the minimum absolute value of f(t) or expressed mathematically.

$$A_c > |\min. f(t)| \tag{2.2}$$

If we consider the case where f(t) is a cosinusoidal signal, that is,

$$f(t) = A_m \cos \omega_m t \tag{2.3}$$

where A_m = amplitude of the cosinusoidal signal

ω_m = angular frequency of the cosinusoidal signal,

then substituting for f(t) in equation (2.1), we obtain:

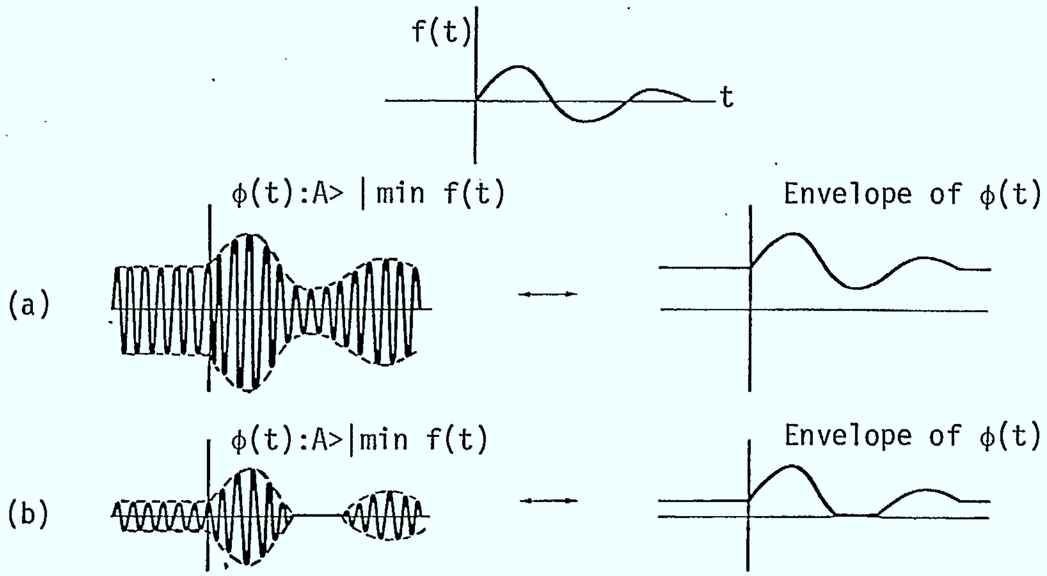


FIGURE 2.2 THE IMPORTANCE OF SUFFICIENT CARRIER LEVEL IN AN AM BROADCAST SIGNAL

$$\begin{aligned}\phi_{AM}(t) &= [A_c + A_m \cos \omega_m t] \cos \omega_c t \\ &= A_c[1 + \frac{A_m}{A_c} \cos \omega_m t] \cos \omega_c t \qquad (2.4) \\ &= A_c[1 + m \cos \omega_m t] \cos \omega_c t\end{aligned}$$

where $m = \frac{A_m}{A_c}$

The dimensionless quantity, m, is generally known as the *modulation index*. When expressed as a percentage it is sometimes referred to as the *modulation depth*. Clearly, m must be less than or equal to 100% if envelope detection is to occur without distortion. If m should exceed 100%, the carrier is then said to be *over modulated*. [In the USA, the FCC permits a positive peak modulation of up to 125% and a negative peak modulation of not greater than 100%; similar regulations are also in effect on an interim basis in Canada].

2.2 Broadcast Spectrum

For ITU Region 2 (North and South America, Greenland), AM broadcasting is allocated the frequency spectrum extending from 535 kHz to 1605 kHz. At present, this band is divided into 107 AM channels, each 10 kHz wide. Carrier frequencies are assigned at 10 kHz intervals from 540 kHz to 1600 kHz (see Figure 2.3).

Since the intermediate frequency (IF) used by a majority of AM broadcast receivers is 455 kHz, the local oscillator (LO) tuning range of these receivers extends from 995 kHz to 2055 kHz. With a 455 kHz IF, the image frequency* band

*For a typical AM broadcast receiver, the IF frequency (f_{IF}) is given by $f_{IF} = f_{LO} - f_s$, where f_{LO} is the local oscillator frequency and f_s, the desired AM carrier frequency. The *image frequency* (f_{image}) is that frequency for which $f_{image} - f_{LO}$ is also equal to f_{IF} or alternatively, $f_{image} = f_s + 2f_{IF}$.

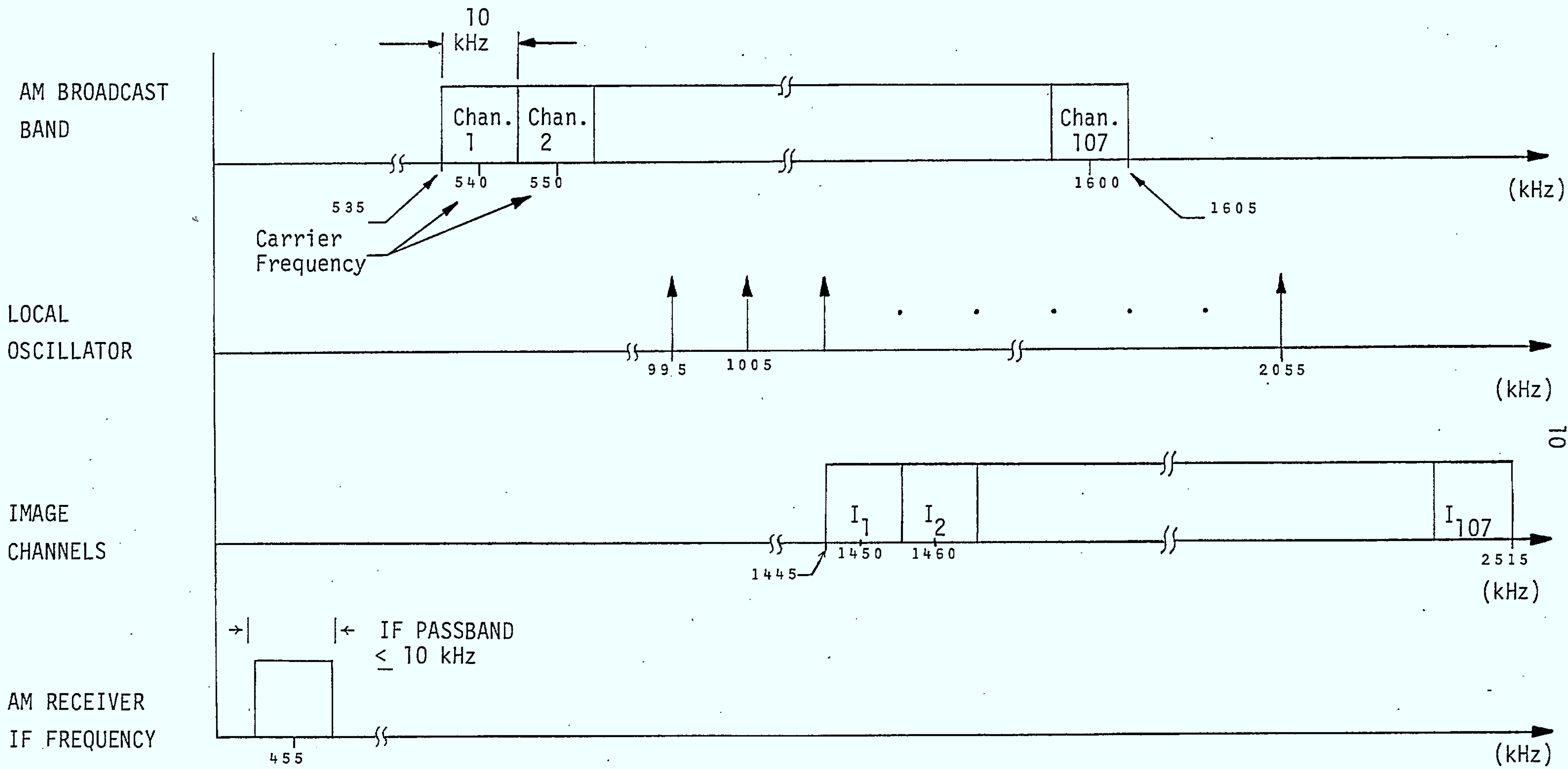


FIGURE 2.3 AM BROADCAST SPECTRUM WITH LO AND IMAGE FREQUENCY RANGES FOR A 455 kHz IF

extends from 1445 kHz to 2515 kHz and clearily overlaps a portion of the AM band. Unless receivers provide sufficient attenuation at the image frequencies, image channel interference will result and this places a constraint on the channels that may be assigned within a given area of broadcast coverage. The latter situation is further complicated by the fact that not all receivers use an IF frequency of 455 kHz and hence the image frequency band will not be the same for all receivers.

In another important class of AM broadcast receivers - the automotive AM receiver - an IF frequency of 262.5 kHz is sometimes used. The LO frequency range, in this case, is therefore 802.5 kHz to 1862.5 kHz and the image frequency range, 1060 kHz to 2130 kHz (Figure 2.4).

Since there are approximately 600 broadcast stations presently sharing the 107 channels available in the AM band, there is an obvious need to utilize this limited resource as efficiently as possible. The use of a reduced carrier spacing of 9 kHz instead of 10 kHz is being contemplated.

An extension of the present AM band is also being considered and submissions have been made by Canada to the ITU at the World Administrative Radio Conference (WARC) held in Geneva (September 1979) for an extension of the AM band from 1605 kHz to 1705 kHz. This submission has been succesful, as discussed in the Introduction (Section 1.1).

The use of the common AM band for the North American Region is governed by agreements established between the following countries: Bahamas, Canada, Cuba, Dominican Republic, Jamaica and the U.S.A. and which are embodied under the North American Regional Broadcasting Agreement (NARBA,

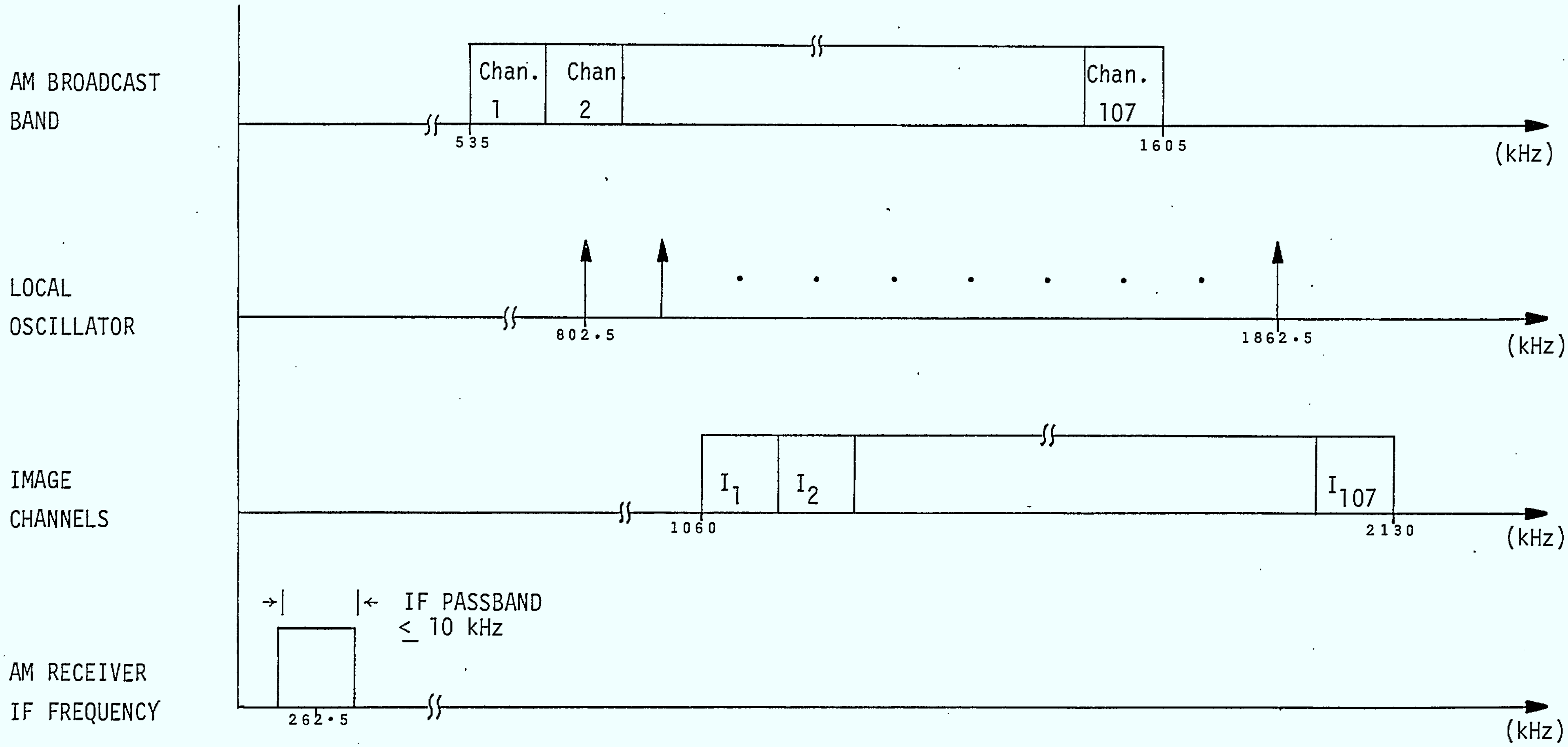


FIGURE 2.4 LOCAL OSCILLATOR AND IMAGE FREQUENCY RANGES FOR A TYPICAL AUTOMOBILE AM RECEIVER

Washington, D.C., 1950).

2.3 Propagation Of AM Signals

During the day, propagation of AM broadcast signals occurs via the ground wave, any sky wave generated being absorbed by the D-layer of the ionosphere. In order to maximize ground wave field strength, the wave must be vertically polarized, that is, the plane of the electric vector must be perpendicular to the earth's surface.

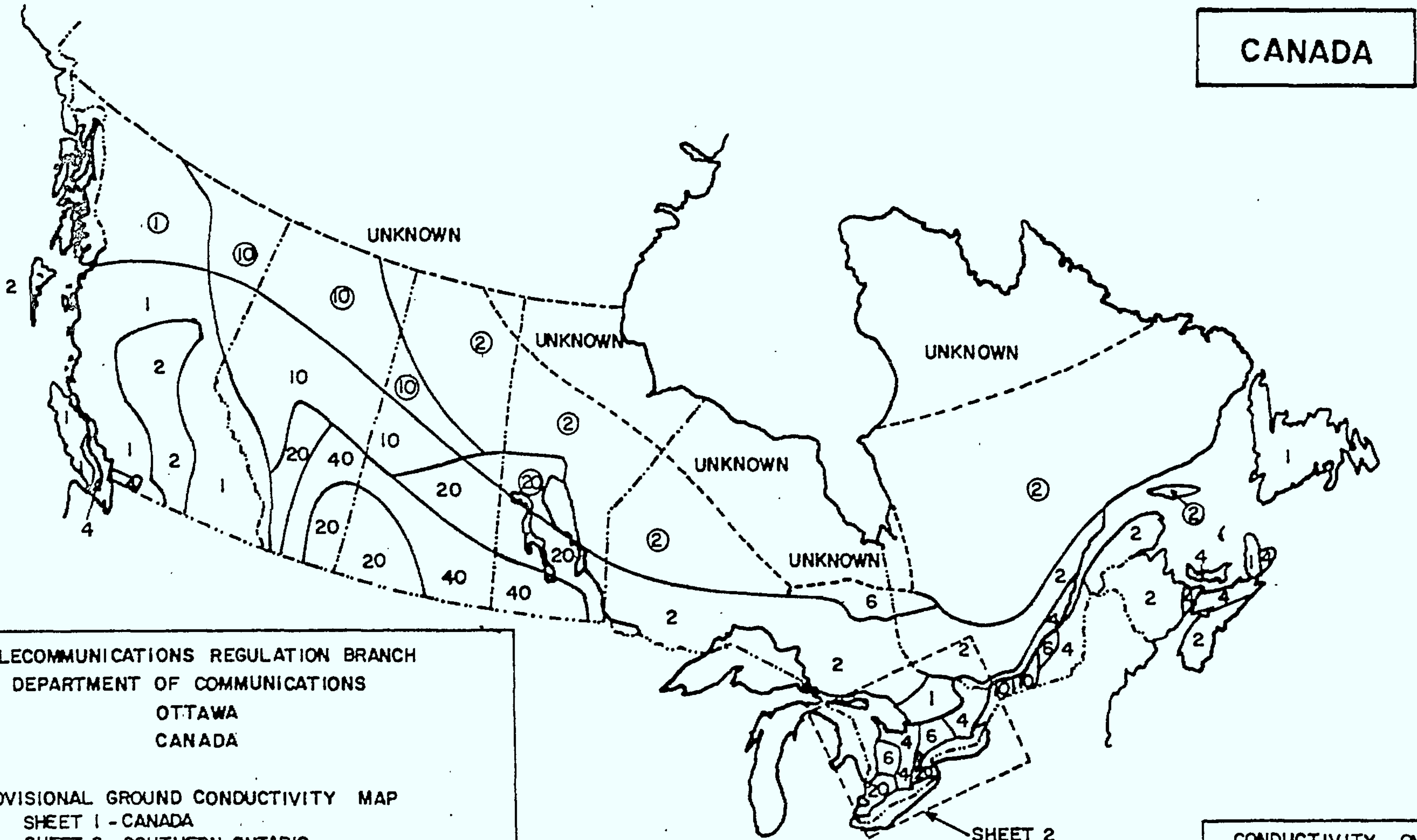
Since the wave induces currents in the ground over which it passes, it thus loses energy due to absorption. Energy loss is greatest over poor-conductivity ground such as granite rocks and least over sea-water where propagation approaches the inverse square law. Ground conductivities* for Canada and Southern Ontario are illustrated in Figures 2.5 and 2.6 respectively.

Energy loss also occurs for a second reason. Since the radius of curvature of the wavefront at MF frequencies (30 kHz to 3000 kHz) is less than the earth's radius, diffraction of the wavefront occurs and increases with distance from the transmitter. This results in increasing attenuation of the electric field component and hence a reduction of electric field strength. It is thus apparent that the field strength at a given AM broadcast receiver is a function of its distance from the transmitter, the power of the transmitter (this may

*Reproduced from Broadcast Procedure 1, Rule 7, issued by the Telecommunications Regulation Branch of the Department of Communications.

FIGURE 2.5

SHEET - I



TELECOMMUNICATIONS REGULATION BRANCH
DEPARTMENT OF COMMUNICATIONS
OTTAWA
CANADA

PROVISIONAL GROUND CONDUCTIVITY MAP
SHEET I - CANADA
SHEET 2 - SOUTHERN ONTARIO
IN THE APPLICABLE AREA, SHEET 2 WILL TAKE PRECEDENCE
WHERE CONDUCTIVITY VALUES ARE NOT CIRCLED, INDICATED VALUES WILL APPLY
WHERE CONDUCTIVITY VALUES ARE CIRCLED INDICATED VALUE IS BASED ON LIMITED DATA ONLY AND SHOULD BE CONFIRMED
WHERE CONDUCTIVITY VALUE IS INDICATED AS UNKNOWN, NO ACCURATE INFORMATION IS AVAILABLE AT PRESENT

JUNE 1, 1960

SPECIMEN ONLY

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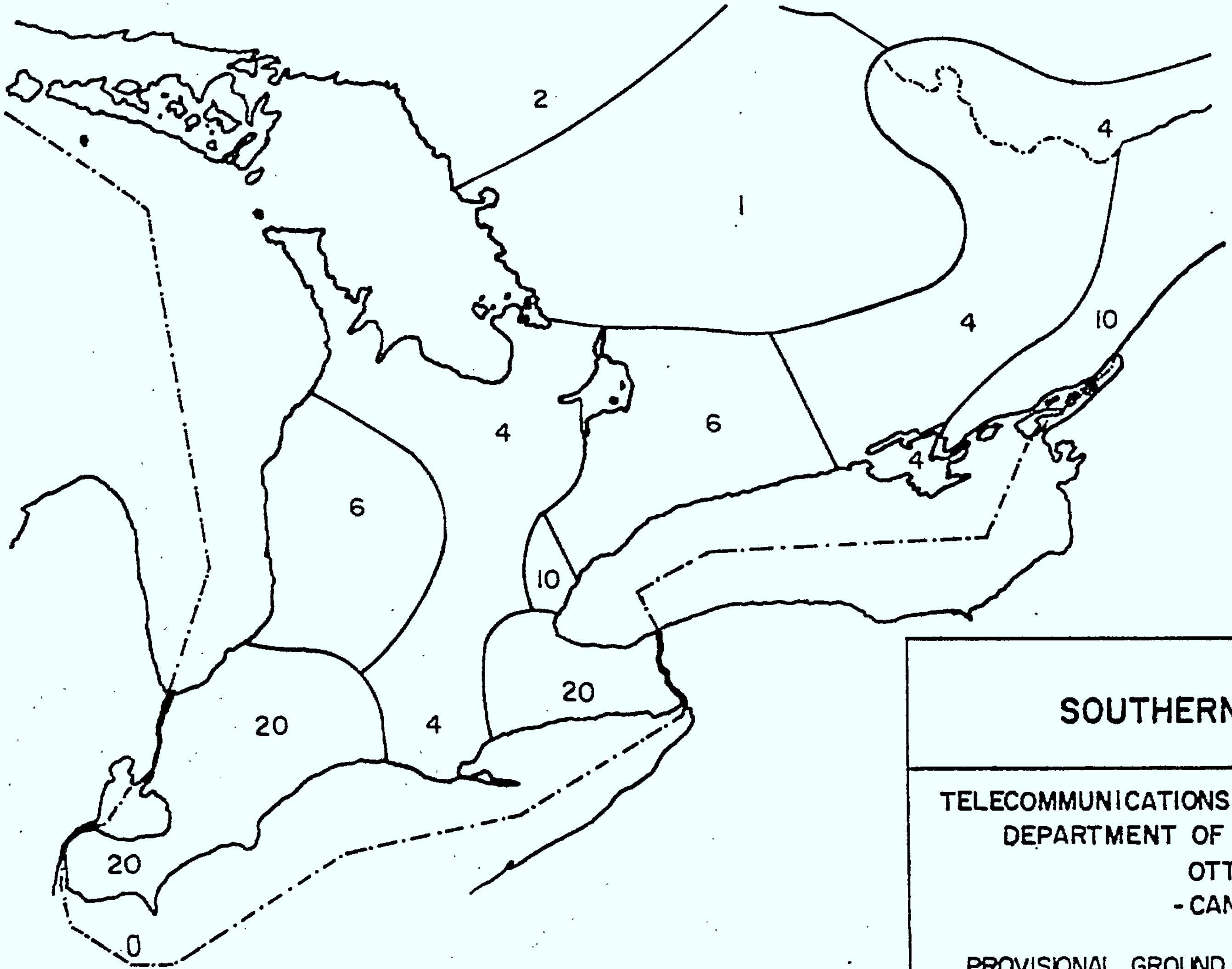
CONDUCTIVITY OVER WATER	
SEA WATER	5000
FRESH WATER LAKES (EXCEPT GREAT LAKES)	10
GREAT LAKES:	
LAKE SUPERIOR	8
LAKE HURON AND GEORGIAN BAY	10
LAKE ERIE	10
LAKE ONTARIO	5

FIGURE 2.6

SHEET - 2



SOUTHERN ONTARIO

TELECOMMUNICATIONS REGULATION BRANCH
DEPARTMENT OF COMMUNICATIONS
OTTAWA
- CANADA -

PROVISIONAL GROUND CONDUCTIVITY MAP

SHEET 1- CANADA
SHEET 2- SOUTHERN ONTARIO

IN THE APPLICABLE AREA SHEET 2 WILL TAKE PRECEDENCE

SPECIMEN ONLY

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vary from a few hundred watts up to 50 kilowatts) and the ground conductivity over the line-of-sight transmission path.

At night, however, refraction of the radiated signal by the ioniospheric E-layer, due to the disappearance of the absorbing D-layer, can result in sporadic sky-wave coverage over thousands of miles and significant co-channel and/or adjacent channel interference to the service areas of distant transmitting stations. Sky-wave propagation is, moreover, subject to considerable fading due to changes in the E-layer and at a distance of 185 miles to 250 miles (300 km to 400 km) from a transmitter, it may produce a field strength comparable to that of the ground wave. Since the phase of the sky-wave is essentially random, the possibility of mutual (destructive) interference betwen the two is then greatest and severe fading can occur. It is understood that such a fading mechanism has indeed been experienced in some locations in the Prairies.

2.4 Service Area

2.4.1 Signal Strength

In establishing the service area of an AM broadcast transmitter, the primary objectives are to provide adequate service to some central population center comprising the "metropolitan area"* and to give maximum coverage to its environs with a minimum of interference to and from other AM broadcast stations which may also be serving the same or nearby areas.

*Broadcast Procedure No.1, Rule 2, defines a metropolitan area as being any area in which there are located, in a reasonably continuous fashion, industrial or residential buildings on parcels of ground known as building lots.

To provide adequate service within a stations *primary service* areas, the following minimum field strengths† are deemed applicable to the areas listed below:

Area	Minimum Field Intensity
(a) Business and/or factory areas of a city	25 to 50 mV/m
(b) Residential	5 to 10 mV/m

The minimum field strength for rural areas is not specified but a value between 0.1 and 1 mV/m is probably adequate for all AM receivers except the portable pocket radios.

These field strength values are established on the basis of the usual noise levels encountered in the respective areas and assume that there is no objectionable limiting interference from other broadcast stations.

A station is deemed to deliver *secondary service* to those areas in which the sky wave has a field intensity of 0.5 mV/m or greater for more than 50% of the time. The secondary service is necessarily subject to interference and fading whereas the primary-service areas of a station are not.

Under the North American Regional Broadcasting Agreement (NARBA), the permissible level of interfering signal at a given field intensity contour of a station is a function of:

(a) the class of the channel being used by the station;

† Broadcast Procedure No. 1, Rule 2.

(b) the class of the station;

(c) whether it is daytime or nighttime transmission;

(d) whether the interference is co-channel or adjacent channel.

There are three classes of channels specified under NARBA:

1. Clear Channel

A clear channel is one to which are assigned primarily one or more Class I stations which are protected from interference so that broadcasting service may be provided over extensive areas by means of ground-wave as well as sky-wave signals.

2. Regional Channel

A regional channel is one to which are assigned several broadcasting stations which are protected from interference so that service may be provided over extensive areas by means of ground-wave signals. No protection from interference is generally provided, however, for service obtained from sky-wave signals transmitted on this type of channel.

3. Local Channel

A local channel is one to which are assigned many broadcasting stations which are protected from interference so that service may be provided over limited areas by means of ground-wave signals. No protection from interference is provided for service obtained from sky-wave signals transmitted on this type of channel.

The channels in the AM band designated under the above categories are listed in Part II of NARBA.

There are basically four classes of stations (Class I to Class IV) with four sub-classifications of Class I stations (Class IA to ID). Class I and II stations are the only types of stations allowed to use the clear channels. A Class II station provides service via ground-wave signals only while a Class I station provides service via both ground-wave and sky-wave signals. Class III stations operate on regional channels only while Class IV stations may use either regional or local channels. Both Classes III and IV stations provide service via ground-wave signals only.

2.4.2 Daytime Interference

For all classes of stations except Class I the permissible co-channel interfering level during daylight hours is 25 µV/m and is determined at the 0.5 mV/m field intensity contour (i.e., a co-channel protection ratio (desired-to-undesired) of 20:1).

In the case of adjacent channel interference, the protection ratio (desired-to-undesired) provided against the first adjacent channel (± 10 kHz) is 1:1 and for the second adjacent channel (± 20 kHz), 1:30.

2.4.3 Nighttime Interference

2.4.3.1 Co-Channel

For broadcasting stations on the same channel, a night-time interference-free service is based on a minimum ratio of 20 to 1 desired ground-wave field intensity to undesired sky-wave field intensity. The procedure for determining the nighttime service area for Classes II, III and IV broadcast stations is outlined in Appendix G of NARBA.

The level of an interfering sky-wave signal is determined by reference to the 10% statistical curve of Appendix E, NARBA. The interference from two or more 10% interfering sky-wave signals to a desired ground-wave signal is taken to be the root-sum-square (RSS) value of the interfering field intensities (except in the case of Class IV stations operating on local channels). Calculation of the RSS value is accomplished by considering the 10% interfering signals in order of decreasing magnitude, adding the squares of these values, and extracting the square-root of the sum, excluding those signals which are less than 50% of the RSS value of the higher signals levels already included.

The RSS value is not considered to be increased when a new interfering signal is added if the level of the new interfering signal is:

(a) less than 50% of the RSS value of the interference from existing stations and,

(b) also less than the smallest signal included in the RSS value of interference from existing stations.

The application of the "50% exclusion" method of calculation may result in anomalies. For example, the addition of a new interfering signal or the increase in value of an existing interfering signal may cause the exclusion of a previously included signal and may cause a decrease in the calculated RSS value of interference. In such cases, an alternate method of calculating the RSS value of interference should be employed.

2.4.3.2 Adjacent Channel

For protection from adjacent channel ground-wave interference,* the nighttime protected ground-wave contour is as follows (Broadcast Procedure No.1, Rule 17):

(a) For stations designated under NARBA as Class I stations, the nighttime protected ground-wave contour is the 0.5 mV/m contour.

(b) For stations designated as Classes II, III or IV, The nighttime protected ground-wave contour of all domestic stations is the 0.5 mV/m contour or 20% of the station's RSS nighttime limitation contour, whichever is greater.

The maximum permissible level of interfering signal on the nighttime protected ground-wave contour of a station is:

Channel Separation Between Stations	Maximum Level of Interfering Ground-Wave Signal
10 kHz	0.5 mV/m
20 kHz	15 mV/m

*Adjacent channel sky-wave interference is generally insignificant and is therefore ignored.

3.0 THE AM BROADCAST RECEIVER

This section examines in some detail the various stages of a typical AM broadcast receiver whose design, almost exclusively, is based on the superheterodyne principle. The objective is to determine what design modifications, if any, may suitably be considered in order to promote more efficient usage of the presently congested AM band and to estimate the possible cost impact of such changes. The audio power amplifier and loudspeaker stage following the baseband detection process will, however, be excluded from these discussions since it has no impact on spectrum utilization.

3.1 CCIR Specifications

In order to foster and promote the large-scale production of low-cost sound-broadcasting receivers so that they might be more available to those areas of the world in which the density of receivers is particularly low, the CCIR has established a set of minimum performance specifications for three types of receivers, namely:

1. Type A Receiver: a low-sensitivity receiver for operation in band 6 (MF - 300 to 3000 kHz).

2. Type B Receiver: a combined receiver for operation in bands 6 (MF) and 7 (HF - 3 to 30 MHz).

3. Type C Receiver: a medium sensitivity frequency modulation (FM) receiver for operation in band 8 (VHF - 30 to 300 MHz).

The only receiver which is of immediate interest here is the Type A or the familiar AM broadcast receiver.

CCIR Recommendation 415[1] - "Specifications for Low-Cost Sound Broadcasting Receivers" - recommends that the minimum performance specifications for the Type A receiver should be as shown in Table 3.1.

3.2 Receiver Classification

AM receiver designs currently fall into three categories:

1. Portable battery-powered receivers without an external power supply
These units vary in size from small pocket radios operating on penlite cells to larger hand-carried units using D cells for power. In price, they vary from approximately $5 (low cost) for the pocket-sized units to $70 (medium cost) or more for the hand-carried units. The audio power output of the pocket receivers is usually in the region of 75 mW while that of the larger hand-carried units is about 250 mW.

2. Console-Type AM receivers powered by an a.c. power line
These units are typically part of an AM/FM Stereo console receiver with high audio output power, i.e., ranging from about 15 watts up to 100 watts or more. This type of AM receiver usually costs more than $100 (premium cost).

[1] CCIR XIIIth Plenary Assembly, Geneva, 1974, Greenbook Vol. X, pp. 119 - 123.

	PARAMETER	SPECIFICATIONS
1.	Frequency Coverage	525 to 1605 kHz
2.	Sensitivity (50 mW output 30% modulation at 400 Hz)	5 mV/m (with a built-in antenna with facilities for using an external antenna)
3.	Signal-to-noise ratio (for input as in 2 above	20 dB mains-operated tube receiver)
		20 dB (transistor receivers)
4.	Power output (< 10% distortion)	Not less than 100 mW
5.	Overall selectivity at - 6 dB points	Passband not less than ± 3 kHz
	at - 20 dB points	Passband not greater than ± 10 kHz
6.	Image, intermediate frequency and spurious response ratio	Not less than 30 dB
7.	Overall fidelity including acoustic response of loud-speaker or,	250 to 3150 Hz, within 18 dB limits
	Alternatively, it may be more convenient for some manufacturers to consider only the electrical characteristics which should be	100 to 4000 Hz within 12 dB limits (in a graphical presentation, 400 Hz should be taken as the reference 0 dB level).

TABLE 3.1 CCIR RECOMMENDATION 415 FOR TYPE "A" (MF) SOUND-BROADCASTING RECEIVERS

3. <u>Automobile AM receivers operated from a 12 volt battery</u> These units may either be strictly AM receivers or part of AM/FM receivers with costs varying from the medium to the premium cost range. The audio power output of these units are relatively high (2 to 3 watts) necessitated by the high ambient noise environment of an automobile.

3.3 Basic Receiver Description

Modern AM broadcast receiver designs are based almost exclusively on the superheterodyne principle in which the incoming modulated carrier (AM) signal is translated in frequency to a lower, fixed frequency, the intermediate frequency (IF). The bandwidth occupied at the IF frequency remains unchanged. The IF frequency used by a majority of AM receivers is 455 kHz but may vary between 455 kHz and 465 kHz. In automobile AM receivers, the IF frequency sometimes used is 262.5 kHz.

The required frequency translation to the intermediate frequency is accomplished by mixing the incoming AM signal with a locally generated signal which differs from the carrier frequency by the I.F. frequency (e.g., 455 kHz). Once at the IF frequency, the received signal can then be easily filtered, amplified and demodulated to recover the baseband modulating signal.

A block diagram of the superheterodyne AM broadcast receiver is illustrated in Figure 3.1. A receiver always has an RF section which is a tuned circuit (which is tunable) connected to the antenna terminals which serves to select the wanted carrier frequency, f_s, and to reject some of the unwanted frequencies. The RF section need not include an RF amplifier and indeed, in many cases, the

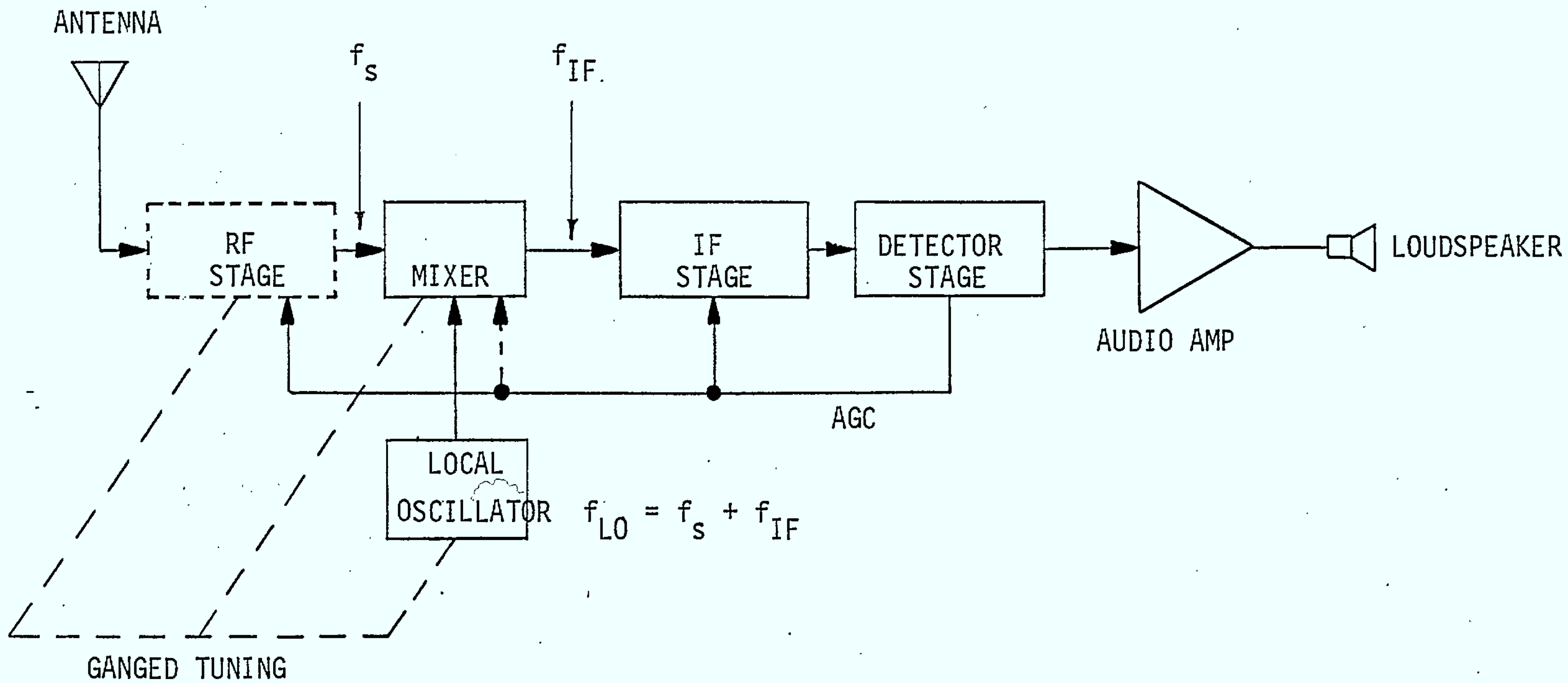


FIGURE 3.1 BLOCK DIAGRAM OF A SUPERHETERODYNE AM BROADCAST RECEIVER

tuned circuit connected to the antenna is the actual input to the mixer circuit.

As mentioned above, the carrier signal is translated to a fixed intermediate frequency, f_{IF}, by mixing it with a local generated signal or local oscillator (LO) whose frequency, f_{LO}, is equal to $f_s + f_{IF}$. The desired signal at the IF frequency (the difference frequency) is then selected from the mixer output by the baseband response of the IF circuitry which is centered on f_{IF}. Since the IF frequency is fixed,it is clearly evident that a constant frequency difference between the RF stage and the local oscillator must be maintained. This is normally achieved by capacitive tuning in which all the capacitors are ganged together and operated in unison by one control knob. In some receivers (mainly the low-cost variety), the local oscillator is combined with the mixer to further simplify the input stages of the receiver.

The IF stage usually consists of two or three amplifier stages which are coupled together by tuned interstage IF transformers. The IF stage provides most of the gain (and therefore sensitivity) required and also determines the noise bandwidth of the receiver.

Almost universally, demodulation of the IF carrier signal is achieved in the detector stage through envelope detection using a diode peak detector. A d.c. voltage, which is proportional to the received carrier level can be simultaneously derived in the peak detector and used for automatic gain control (AGC) of the IF and RF (if provided) amplifier stages (and possibly the mixer) to maintain a fairly constant IF level to the detector under fluctuating carrier level conditions.

Further amplification of the recovered audio signal is provided by the audio power amplifier which is then used to drive a loudspeaker.

A typical gain distribution for an AM broadcast receiver is a follows: 10 to 30 dB in the RF stage, 60 dB in the IF amplifiers and 50 to 60 dB in the audio stages, the total power gain being 100 to 150 dB.

Examples of Commercial AM Broadcast Receivers

Figure 3.2 illustrates a typical commercial 6 - transistor AM broadcast receiver. Note that there is no RF amplifier stage; instead, the self-oscillating mixer stage of Q1 serves as the input stage of the receiver. As a consequence, AGC cannot be applied to the self-oscillating mixer otherwise the local oscillator frequency would vary with AGC bias. AGC is applied in this case to the first IF amplifier stage of Q2 only; none is applied to the second IF amplifier (Q3) because of its high operating level. Consequently, fixed bias for class A operation at all times is provided to prevent overloading and distortion in Q3.

The gain distribution for the receiver is as follows:

Mixer conversion gain	30 dB
IF Amplifiers	50 dB
Diode detection loss	-30 dB
Audio Driver plus output Stage	60 dB
Overall Receiver gain	≃ 110 dB

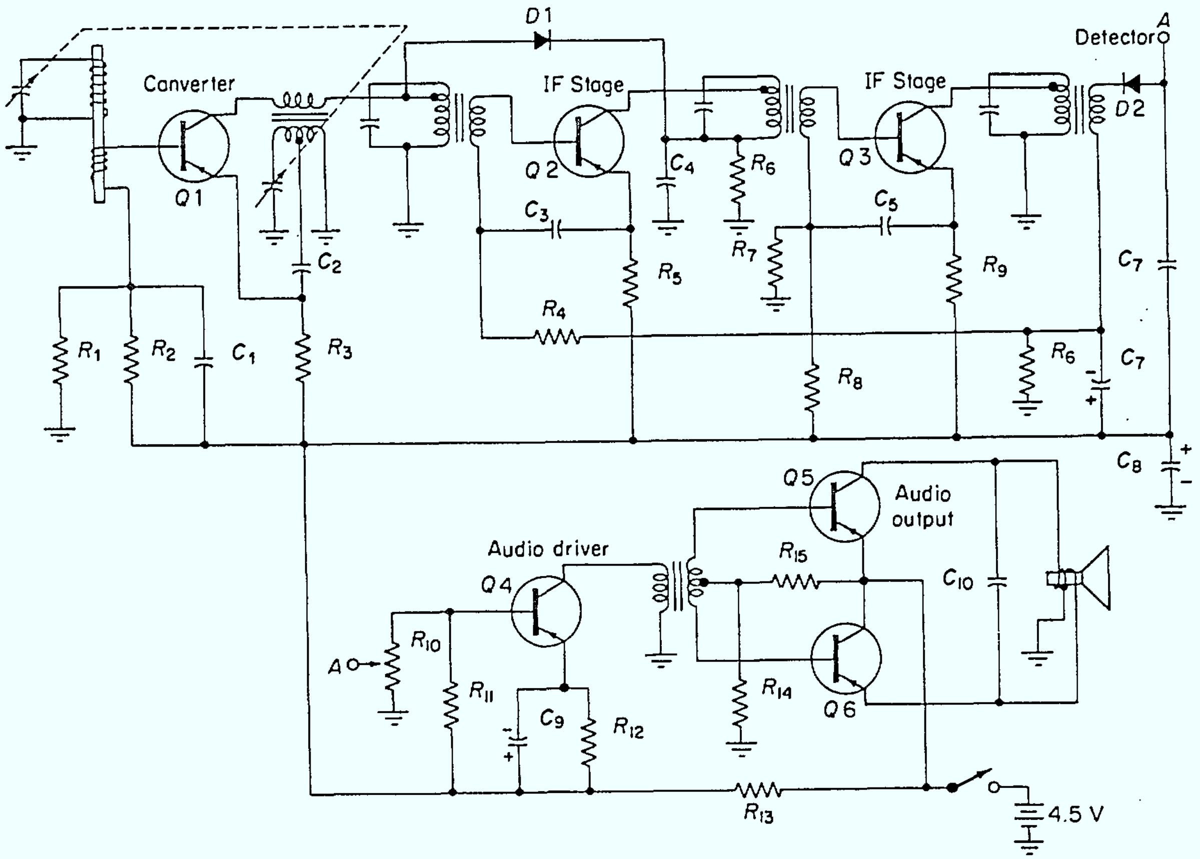


FIGURE 3.2 A TYPICAL 6-TRANSISTOR AM BROADCAST RECEIVER

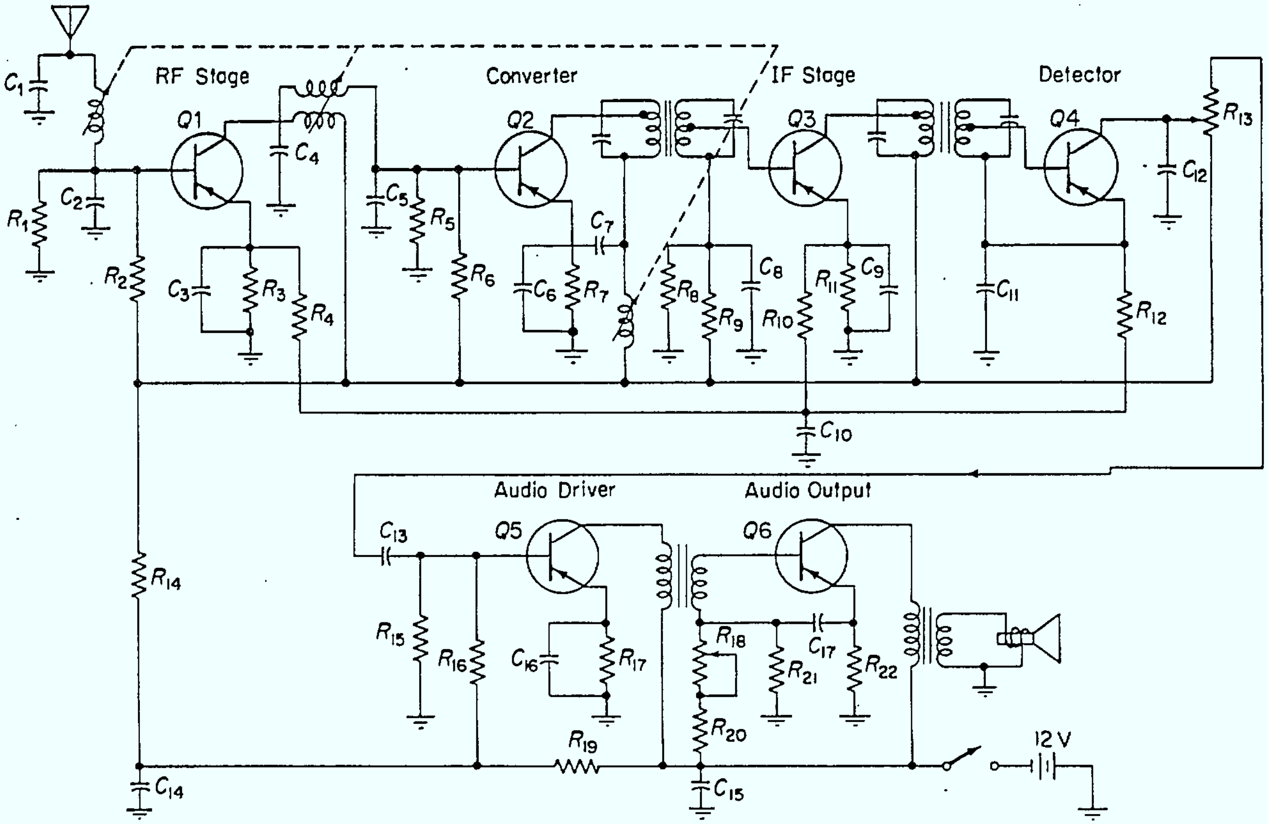


FIGURE 3.3 A MORE ELABORATE RECEIVER DESIGN TYPICAL OF AUTOMOTIVE AM RECEIVERS

Figure 3.3 illustrates a more elaborate receiver configuration which is typical of automobile radio design. An RF amplifier is provided and the audio output stage supplies 4 watts to the speaker. Note that a transistor detector, which provides a gain of 5 dB, is used instead of the diode detector. The gain of the audio driver and output stages is about 65 dB while the IF amplifier has a maximum gain of 30 dB. The gain of the RF and mixer stages is approximately 55 dB thus providing a total available receiver gain of about 155 dB.

In the receiver configuration of Figure 3.3., AGC is applied to both the RF and IF amplifiers. Most of the AGC control is applied to the RF stage and during strong signal receptions, the RF stage may be cut off completely. There is still, however, a small amount of signal which will leak through from the input to the output of Q2 as a result of the junction capacitance of the transistor.

A more detailed examination of the design considerations associated with each stage of the superheterodyne AM receiver (except the audio amplifier stage) follows.

3.4 RF Stage

3.4.1 Antenna

The required circuit sensitivity of a receiver is controlled by the efficiency of the antenna system, given that the receiver must be capable of producing a usable output at some minimum field strength.

There are basically two types of antennas used by AM receivers:

(a) The ferrite rod antenna, which is used in portable and console-type receivers and ,

(b) the capacitive whip antenna, which is used by automative receivers.

Ferrite Rod Antenna

This type of antenna consists of a short piece of ferrite rod on which are wound several turns of wire. The equivalent circuit of the ferrite rod antenna is illustrated in Figure 3.4 where:

L = antenna inductance
C = tuning capacitance plus stray capacitance (typically, 20 - 150 picoFarads)
N_o = antenna turns ratio (primary to secondary)
R_{IN} = receiver circuit input impedance
R_p = equivalent parallel loss resistance
R_L = equivalent loading resistance
V_{IN} = voltage applied to the receiver's circuit
V_{ID} = voltage induced in the antenna
V_T = voltage transferred across the tank circuit.

The unloaded Q of a typical ferrite rod antenna is of the order of 200 and the input voltage to the receiver's circuit, V_{IN}, is given by:

$$V_{IN} = \frac{V_T}{N_o} = \frac{Q_L V_{ID}}{N_o} \doteq \frac{Q_L}{N_o} \ell_{eff} E \qquad (3.1)$$

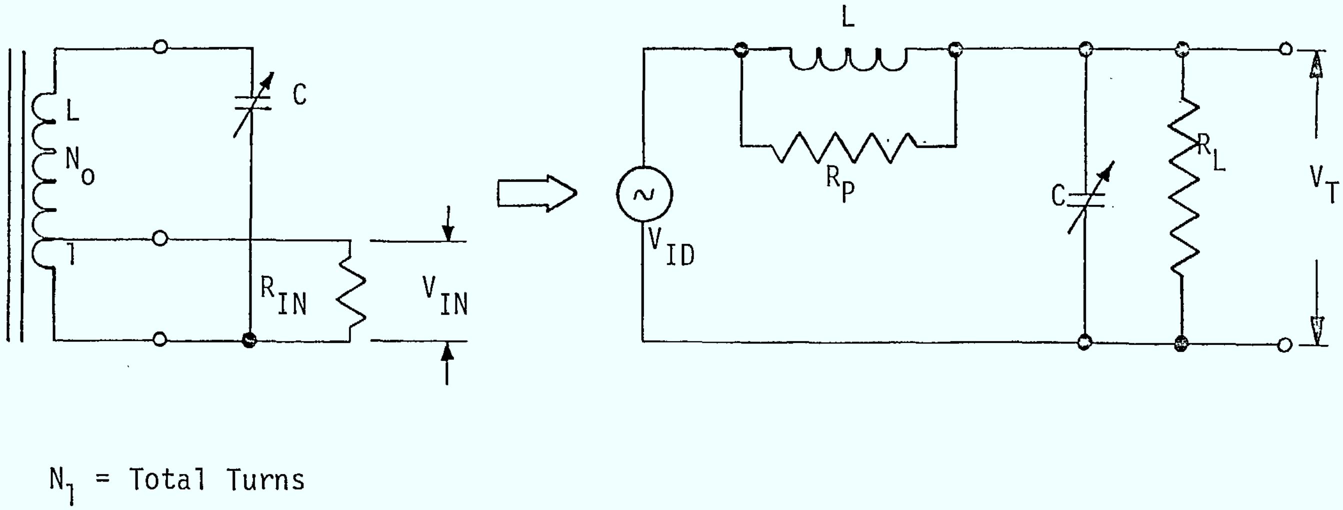


FIGURE 3.4 EQUIVALENT CIRCUIT OF A FERRITE ROD ANTENNA

Q_L = loaded Q of the antenna circuit
E = field strength, in volts/meter
ℓ_{eff} = effective length of the antenna, in meters

Capacitive Whip Antenna

An automotive receiver's capacitive antenna, whose length is adjustable to about 1 meter, can be analyzed in a similar manner to the ferrite rod antenna. Figure 3.5 illustrates the equivalent circuit of such an antenna. C1 is the capacitance of the vertical rod with respect to the horizontal ground plane, while C2 is the capacitance of the coaxial cable connecting the antenna to the automotive AM receiver. In order to obtain a useful signal output, this capacitance is tuned out with the inductor L. Losses in the inductor and the input resistance of the receiver circuitry form R_L.

The voltage transferred across the tank circuit, V_T, is given by:

$$V_T = V_{ID}\, Q_L \frac{C1}{C1 + C2} \qquad (3.2)$$

where, as before:

V_{ID} = voltage induced in the antenna
$= \ell_{eff}$ E volts

ℓ_{eff} = effective height of the antenna, in meters

E = field strength, in Volts/meter

Q_L = loaded Q of the antenna circuit

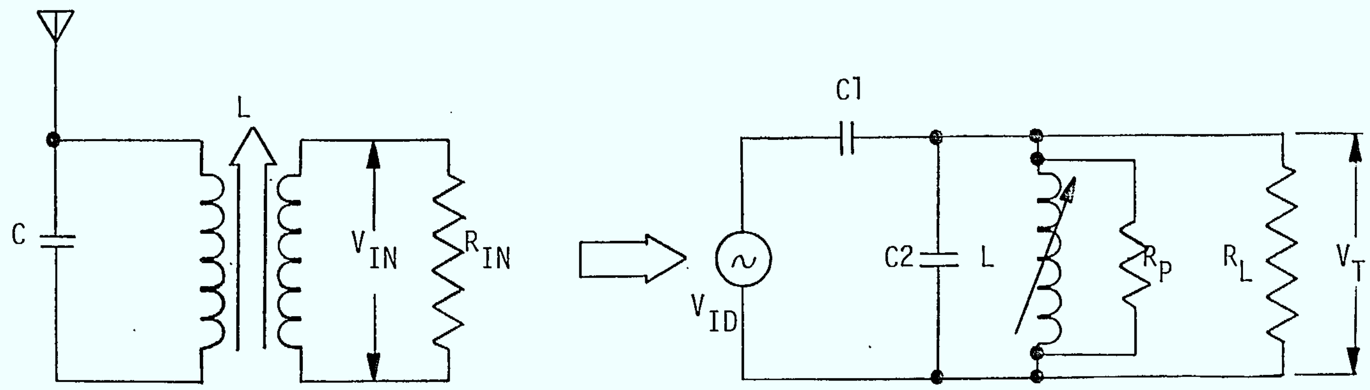


FIGURE 3.5 EQUIVALENT CIRCUIT OF A CAPACITIVE WHIP ANTENNA

Laurent and Carvalho [2] have shown that the effective height of a vertical capacitive whip antenna can be approximated by

$$\ell_{eff} \simeq \frac{h}{2} \tag{3.3}$$

where h = height of the antenna, in meters

Thus substituting for V_{ID} and ℓ_{eff} in equation 3.2, we obtain:

$$V_T \simeq \frac{Q_L\, h\, E\, C1}{2(C1 + C2)} \tag{3.4}$$

Finally, the voltage at the input to the receiver's circuitry, V_{IN}, is therefore:

$$V_{IN} \simeq \frac{V_T}{N_o} \simeq \frac{Q_L\, h\, E\, C1}{2N_o(C1 + C2)} \tag{3.5}$$

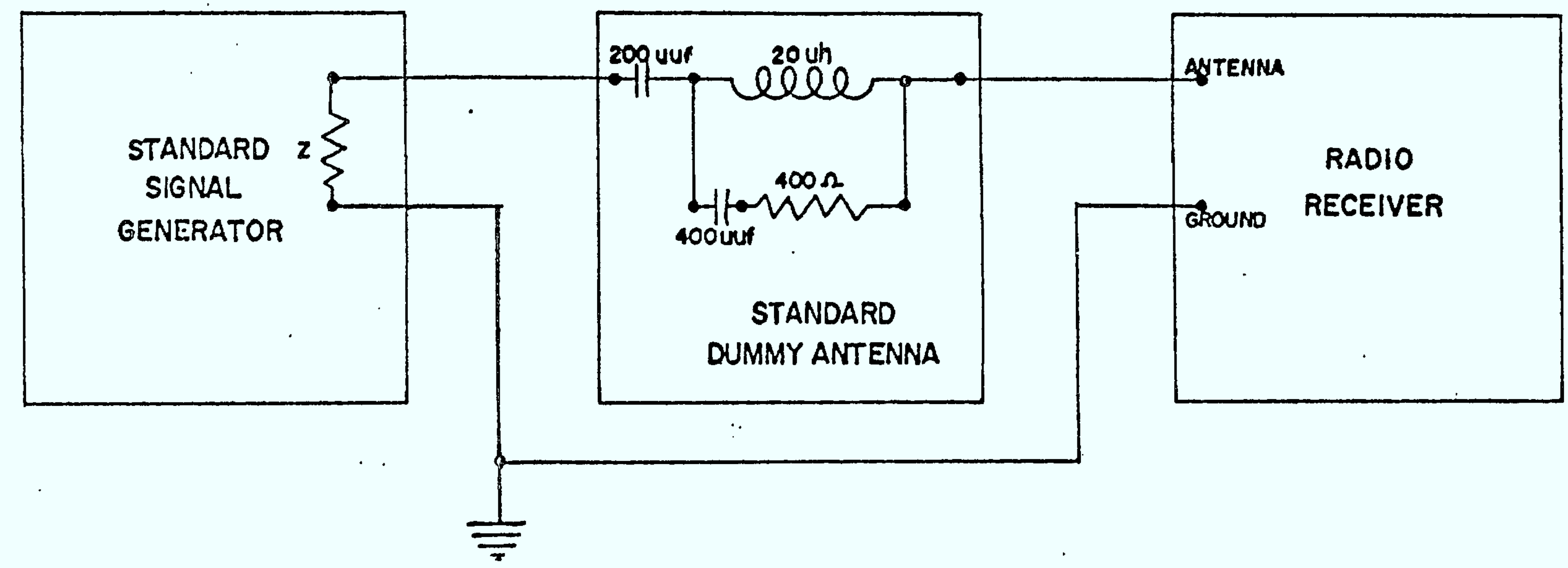
where

$$N_o = \sqrt{\frac{R_L}{R_{IN}}}$$

3.4.2 Sensitivity

The sensitivity of a receiver is basically its ability to amplify weak signals. It is defined in terms of the least signal voltage that must be applied to the receiver's input terminals to give a standard output power at a given signal-plus-noise to noise ratio (S+N/N). For AM broadcast receivers, several of the relevant test parameters have been standardized. For example, a modulation

[2] H.J. Laurent and C.A.B. Carvalho, "Ferrite Antennas for AM Broadcast Receivers", Application Note, Bendix Radio Division of the Bendix Corporation.



NOTE - GENERATOR IMPEDANCE MUST BE SMALL COMPARED TO 400 OHMS

FIGURE 3.6:

STANDARD DUMMY ANTENNA AND METHOD OF CONNECTIONS

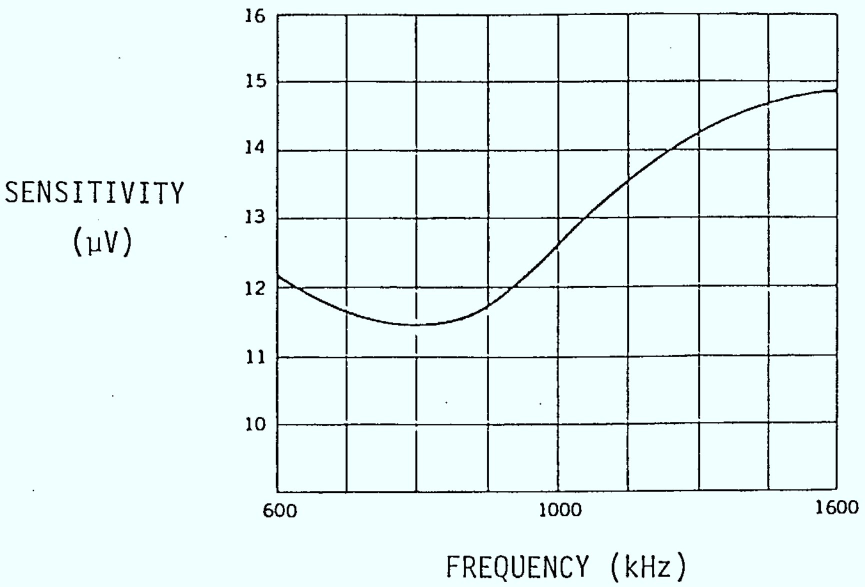


FIGURE 3.7 SENSITIVITY CURVE FOR A GOOD DOMESTIC RECEIVER

depth of 30% and a 400 Hz sinusoidal modulating signal are used. The modulated carrier signal is coupled to the receiver through a standard coupling network known as a dummy antenna (see, for example, Figure 3.6). The standard power output is 50 mW measured into a standard dummy load consisting of a pure resistance whose value is equal to the 400 Hz impedance of the loudspeaker normally used in the receiver.

The most important factors determining the sensitivity of an AM broadcast receiver are the gain of the IF amplifiers and that of the RF amplifier (if there is one). Sensitivity is normally expressed in microvolts or in decibels below 1 Volt and is measured at three points along the tuning range. A typical sensitivity curve for a good domestic or automobile radio is illustrated in Figure 3.7. Portable and other small receivers used only for the AM broadcast band might have a sensitivity in the region of 150 μV.

3.4.3 Selectivity

The selectivity of a receiver is a measure of its ability to reject (adjacent) unwanted signals. It is expressed as a curve, as, for example, in Figure 3.8, which shows the extent to which the receiver attenuates signals at frequencies which are adjacent to the one to which it is tuned. In addition to adjacent channel rejection, a receiver must also provide sufficient rejection against image and direct IF interference.

Selectivity varies with receiving frequency and becomes somewhat worse when the received frequency is raised. The primary adjacent channel selectivity is provided by

(Receiver tuned to 950 kHz)

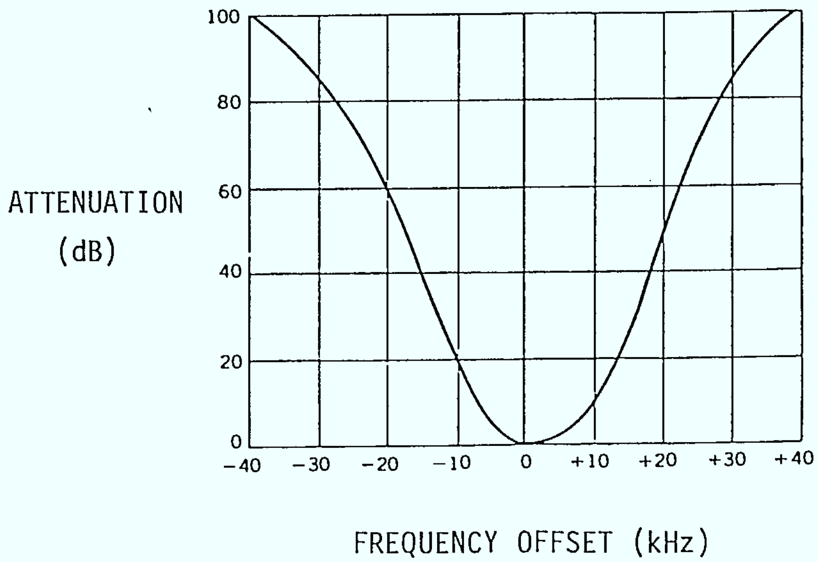


FIGURE 3.8 , TYPICAL SELECTIVITY CURVE FOR AN AM BROADCAST RECEIVER

the IF stages whereas image and direct IF selectivity must be provided by the RF circuitry. In receivers using a ferrite rod antenna and no RF amplifier stage, the RF selectivity is provided entirely by the antenna tuned circuit. A high Q in the antenna coil thus not only provides adequate desired signal level to override the mixer noise but also protects against image and IF interference.

For a single-tuned circuit, the image rejection provided is given by:

$$\text{Attenuation (dB)} = 10 \log_{10}\left[1+\left(\frac{f_{si}}{f_s} - \frac{f_s}{f_{si}}\right)^2 Q_L^2\right] \quad (3.6)$$

where

f_s = frequency of the wanted signal

f_{si} = image frequency = $f_s + 2f_{IF}$

Q_L = loaded Q of the tuned (antenna) circuit.

If there are two tuned circuits, both tuned to f_s, the rejection of each is calculated using equation 3.6 and the total rejection is then (ideally) the sum of the two (in decibels). This situation refers to a receiver which has an RF amplifier stage. An example is the automotive receiver which must provide greater selectivity in order to cope with the strong signals encountered and the extreme dynamic range of signal levels. The image rejection of these receivers at 1400 kHz is typically 58 dB in spite of the lower IF frequency (usually 262.5 kHz), the direct IF rejection at 600 kHz is typically 50 dB and the adjacent-channel selectivity is about 20 dB.

3.4.4 Gain Compression

If the input power versus output power response of an amplifier is plotted on a log-log scale, it will have a 1:1 slope in the linear region of operation. At higher input power levels, however, the amplifier will be driven into saturation (a nonlinear region of operation). In this region of operation, gain compression occurs, i.e., the output power no longer bears a linear relationship with the input power. This effect is exhibited on the log-log plot as a gradual flattening of the response curve (see Figure 3.9). The point at which the actual response deviates from an extrapolation of the linear response by 1 dB is commonly called the 1 *dB compression point*, which provides a performance index of the large-signal handling capability of an amplifier.

Due to the non-ideal linearity of an amplifier, undesired higher order harmonics of the fundamental input frequency are inevitably generated even when the amplifier is operating in its 'linear' region. This results in a fairly small amount of harmonic distortion of the output but becomes more pronounced as the operating point approaches the highly nonlinear region of saturated operation due to the increasing magnitude of the harmonics. The chief contributors to harmonic distortion are the second and third order harmonics since the amplitudes of higher order harmonics decrease rapidly beyond the third harmonic. An amplifier is therefore generally operated at a point of minimum harmonic distortion rather than maximum output.

For AM receivers fitted with an RF amplifier input, it is important to maintain the operating point in the linear region over a wide dynamic range of desired input carrier

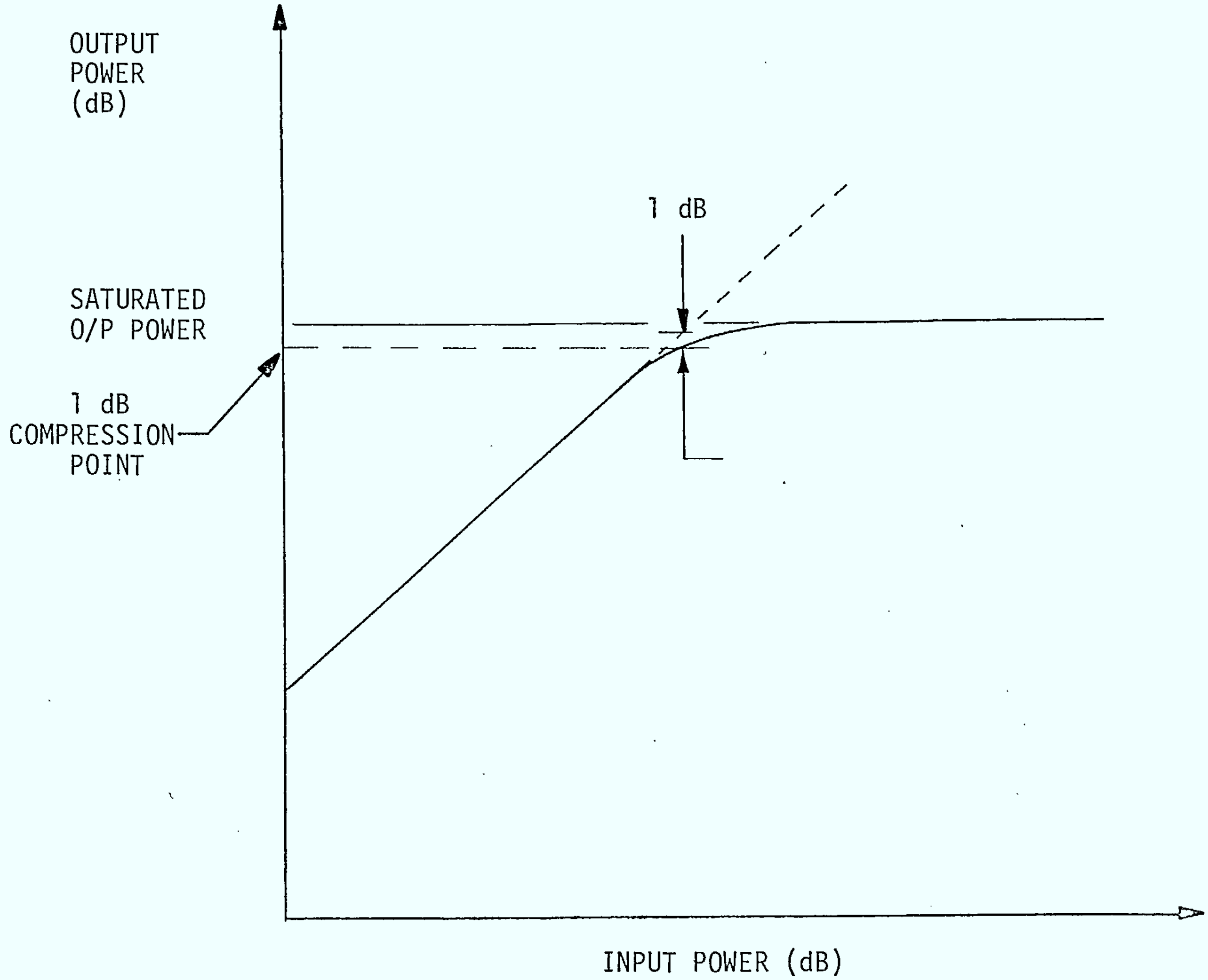


FIGURE 3.9 TYPICAL INPUT/OUTPUT POWER RESPONSE OF AN AMPLIFIER

level for basically two reasons:

1. To prevent possible saturation of either the mixer or the succeeding IF amplifier stages which would result in distortion of the recovered baseband modulating signal.
2. To minimize the level of undesired intermodulation (IM) products resulting from the amplification of two or more input RF signals. (This subject is discussed in the next section).

To prevent saturated operation, automatic gain control (AGC) or delayed AGC (refer to Section 3.6.4) is often applied to a selected number of RF, mixer and IF stages. In some cases, the RF and mixer stages are excluded from the AGC control loop which renders the RF amplifier susceptible to saturation. From the previous discussions, it is apparent that where AGC is not applied to the RF stage, the 1 dB compression point of the amplifier should be designed to be as high as possible. For this purpose, the insulated-gate MOS Field Effect Transistor (MOSFET) provides an excellent alternative to the bipolar transistor by virtue of its high dynamic range, comparatively high linearity of its transfer curves, high input impedance and low noise performance at virtually the same cost.

For mixer applications, the dual insulated gate MOSFET provides a wide dynamic range, excellent isolation between the local oscillator source and the RF signal and reduces LO feedthrough to the antenna.

In most inexpensive receivers, there is no RF amplifier provided. Instead, the local oscillator is often combined with the mixer to form what is commonly known as a self-oscillating mixer. The disadvantage with this approach is that AGC cannot be applied for the reason that the stage biasing and hence the frequency of oscillation would be altered. In some cases, this has resulted in the use of a separate mixer stage to which AGC can be applied.

3.4.5 Intermodulation

When two (or more) signals are amplified in a nonlinear device, spurious products are generated in the output of the device that are harmonically related to the frequencies of the input signals. These spurii are commonly referred to as *intermodulation* or *IM products*. Generally speaking, only the second and third-order IM products are usually of concern because of their much larger magnitudes as compared to the higher order products. In many cases, third-order products present serious concerns because of their tendency to fall within the passband of even moderate-bandwidth amplifiers.

To characterize the IM performance of an amplifier, the input power versus output power response for the third-order IM products is usually plotted on the same graph as the fundamental. On a log-log scale, the third-order response curve is linear below the 1 dB compression point and has a slope of 3:1. A similar plot for the second-order IM response yields a linear response with a slope of 2:1.

The point at which the extensions of the first and third-order responses intersect on the output power scale is generally defined as the *third-order intercept point* or simply, *intercept point* (Figure 3.10). An extension of

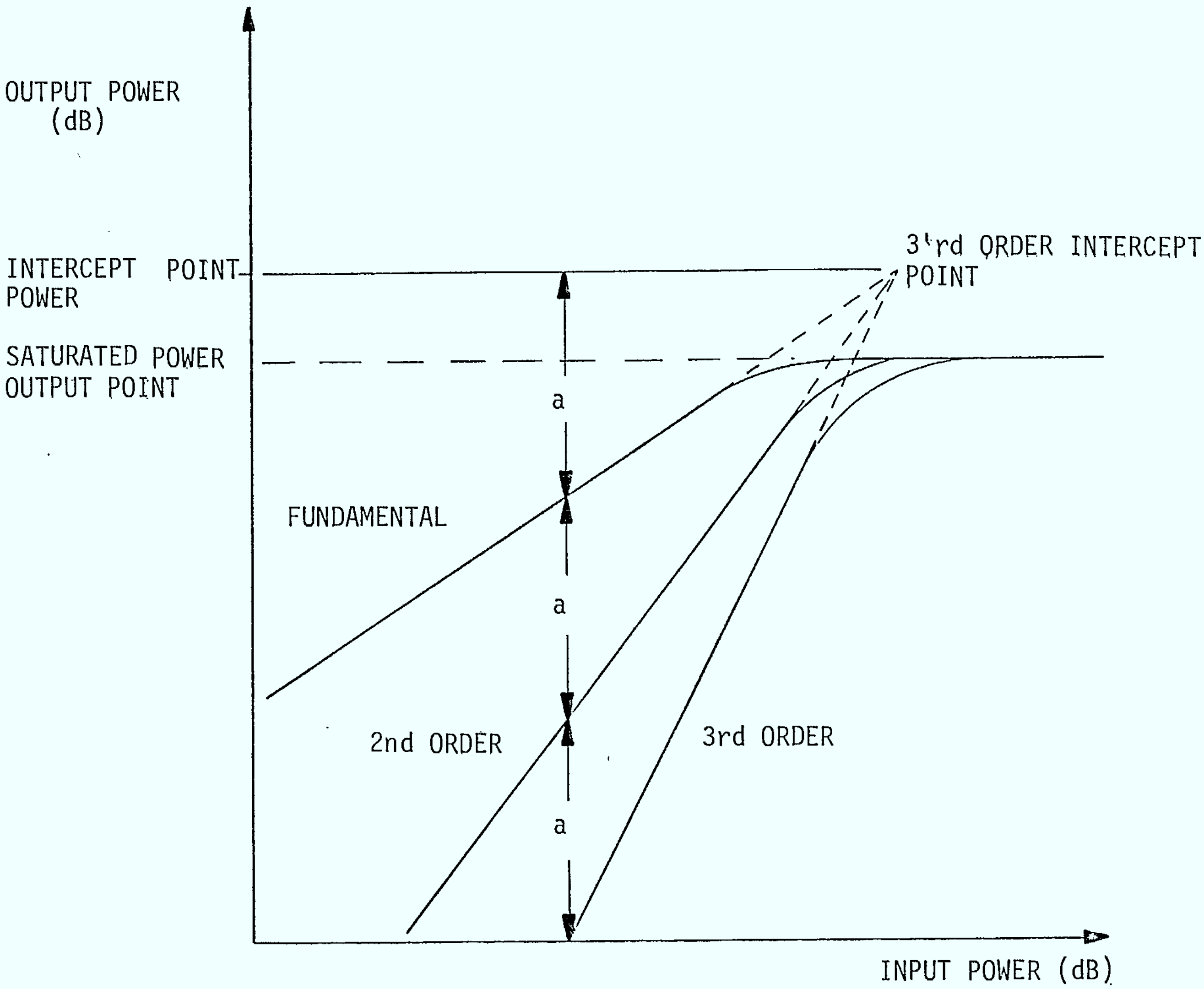


FIGURE 3.10 INTERCEPT POINT RESPONSE OF A TYPICAL AMPLIFIER

the second-order response generally intersects near the same point as well.

The usefulness of the intercept point concept stems from the fact that since the difference between the slope of the third-order curve and the fundamental is 2:1, the third-order IM products will be twice the distance down from the fundamental as the the fundamental is from the intercept point at any output power in the linear range. Similarly, since the difference between the slope of the second-order response curve and the fundamental is 1:1, the second-order IM products will be the same distance down from the fundamental as the fundamental is from the intercept point at any output power in the linear range.

It is evident from Figure 3.10 that, for a given intercept point, minimizing the required power output of the fundamental is the key to reducing the levels of IM products. For receivers fitted with an RF stage, the gain of the antenna plus amplifier is thus generally limited to about 30 dB or less.

3.5 Mixer/Local Oscillator Stage

3.5.1 General Description

The purpose of the mixer or converter stage is to translate the desired AM broadcast signal from a carrier frequency in the AM band (540 kHz to 1600 kHz) to a fixed intermediate frequency (IF) of 455 kHz (typically). Two basic conditions are required for frequency conversion to occur:

1. A local oscillator (LO) signal must be provided and,

2. A nonlinear device must be used for mixing.

The frequency of the LO signal is chosen to be equal to the sum of the desired frequency and the IF frequency for reasons which will be explained later in this section.

The output frequency spectrum of the mixer will contain the sum and difference frequencies - the difference frequency being the desired IF - plus higher order harmonics and intermodulation products of the two fundamental input frequencies. The desired IF frequency is then extracted by the bandpass response of the IF stage which rejects all other extraneous frequencies that are outside its passband.

The mixer may be located either immediately following the antenna tuned circuit or following the RF amplifier, if one is utilized. The mixer circuit may either be a heterodyne mixer with a separate local oscillator or it may be a self-oscillating mixer arrangement.

The nonlinear device commonly used in AM receivers for mixing is a transistor which has been approximately biased to utilize the nonlinearity of its collector current versus base-emitter voltage characteristics.

A key requirement of the mixer/LO arrangement is the accuracy with which the LO can be tuned in conjunction with the antenna tuned circuit to maintain a relatively fixed IF frequency output over the entire AM band.

3.5.2 Circuit Types

There are basically three types of mixer/LO (or tuner) circuits used in AM receivers. The simplest and most economical of these is the self-oscillating mixer circuit which is commonly used in inexpensive receivers for obvious reasons. Figure 3.11 illustrates this type of mixer. A drawback of this mixer circuit is that AGC cannot be applied to the circuit since a change in the biasing conditions would result in an alteration of the local oscillator frequency. The self-oscillating mixer is hence susceptible to overload under strong signal conditions.

The second type of mixer overcomes this disadvantage by separating the mixer and the local oscillator circuits. By doing so, AGC may then be applied to the mixer circuit alone without affecting the local oscillator. This configuration is depicted in Figure 3.12. Note that in this particular circuit there is no RF amplifier provided.

The third type of mixer circuit is, in fact, simply a variation of the second type in which inductor tuning

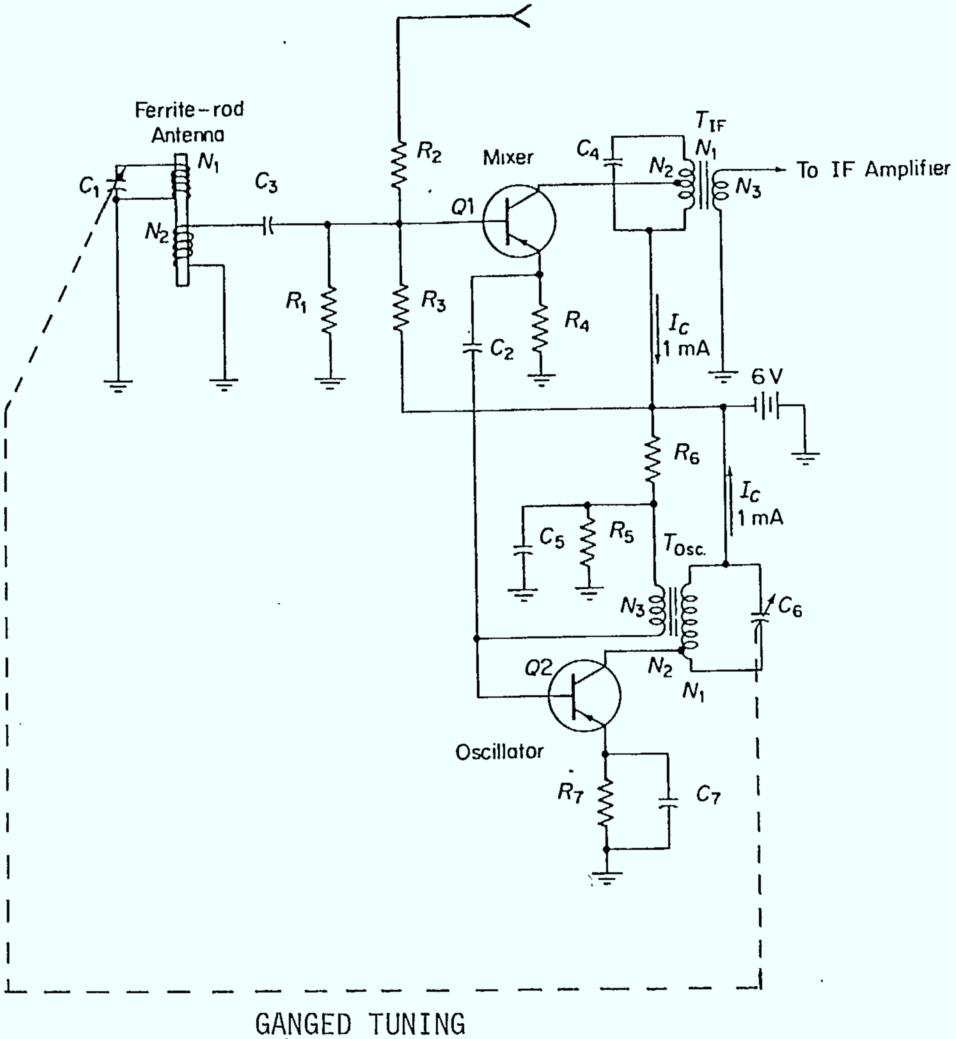


FIGURE 3.11 SEPARATE MIXER AND LOCAL OSCILLATOR ARRANGEMENT

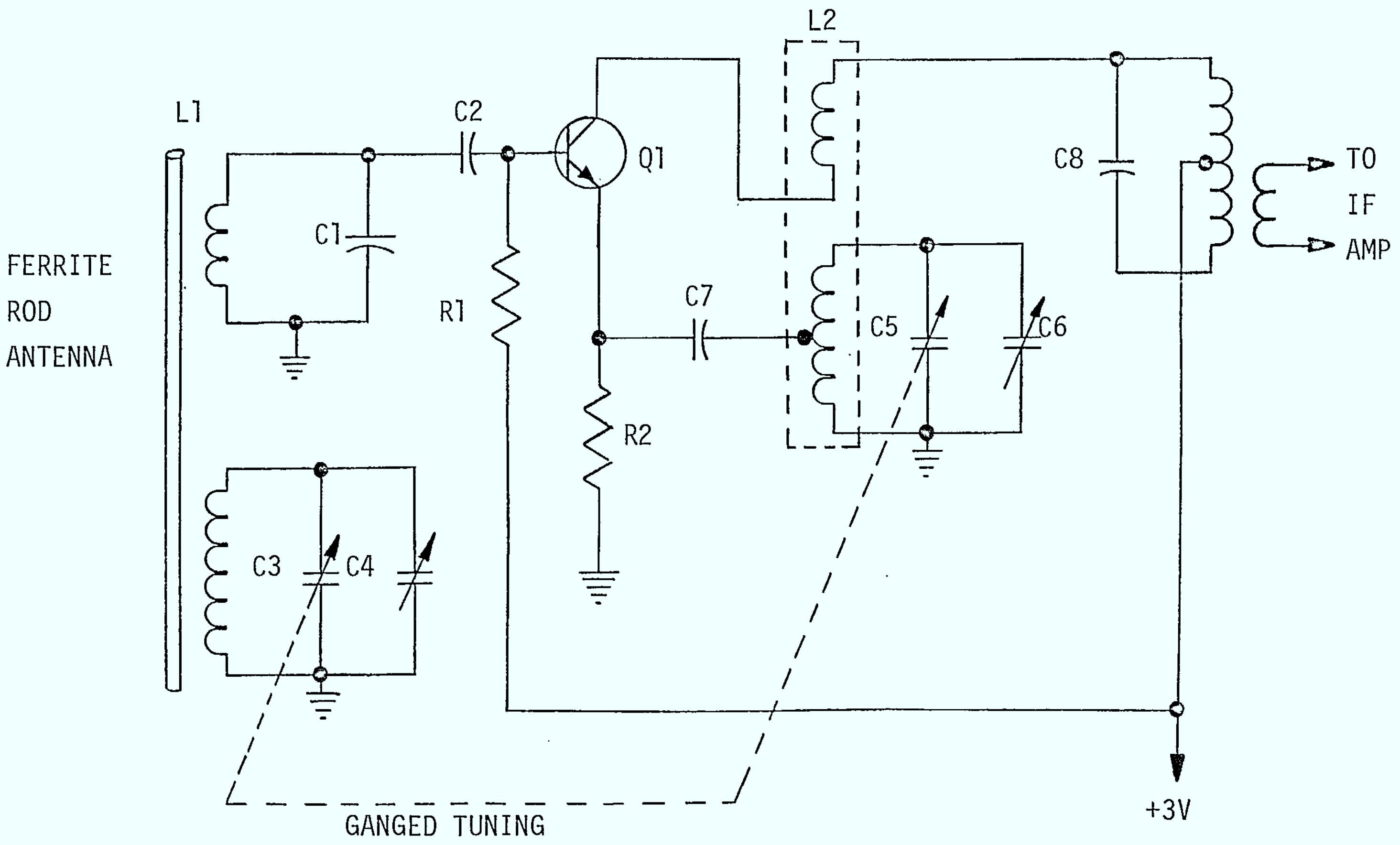


FIGURE 3.12 SELF-OSCILLATING MIXER CIRCUIT

replaces capacitor tuning. It is found mainly in automotive receivers which rely on push-button tuners and slide mechanisms to adjust core positions in coils rather than resetting the exact angle on the rotating shaft of a variable capacitor. An example of this type of tuning arrangement was previously presented in Figure 3.3.

3.5.3 Tuning Range

The tuning range of an AM receiver's local oscillator is determined on the basis of a carrier frequency range of 540 kHz to 1600 kHz and an IF frequency which is very often 455 kHz. For the usual case of LO frequency above signal frequency, this range is 995 kHz to 2055 kHz, giving a ratio of maximum-to-minimum frequency of 2.1:1. If the LO frequency had been chosen to be below the signal frequency, the LO range would have been 85 kHz to 1,145 kHz and the ratio 13.5:1. The normal tunable capacitor has a capacitance ratio of 10:1 which gives a frequency ratio of 3.2:1. Clearly the 2.1:1 ratio required of a local oscillator operating above the signal frequency can be accomodated by the capacitor tuning range whereas the other system has a frequency range which cannot be covered in a single sweep. This is the basis for choosing the LO frequency to be higher than the desired signal frequency.

A further point to note is that the antenna tuned circuit covers the range from 540 kHz to 1600 kHz, i.e., a maximum-to-minimum frequency ratio of 3:1, whereas the LO's frequency ratio is 2.1:1. Since these ratios are not equal but the oscillator and antenna tuning capacitors are ganged together on the same control, the capacitor plates for the oscillator must be shaped to provide proper tracking.

Tuning Range Extension

With the proposed extension of the upper limit of the AM band from 1600 kHz to 1700 kHz, the maximum-to-minimum LO frequency ratio would then be 2.2:1 or a percentage change of 4.9%. The corresponding percentage change in capacitance ratio would be 10%, i.e., from 4.26:1 to 4.69:1, which is clearly well within the 10:1 capability of the present variable capacitors.

In AM receivers, the tuning range of the variable capacitor (typically 5 - 350 pF) is reduced to the required range by means of a "padding" capacitor (discussed in the next sub-section). A reduction of 10% in the capacitance value at the high end of the AM band should therefore present little difficulty in the design of new receivers and would only require minor adjustments in present AM receivers.

3.5.4 Frequency Tuning And Tracking

As illustrated in Figures 3.2 and 3.3, an AM receiver has a number of tunable circuits which must all be tuned correctly if any given station is to be received. For obvious reasons, the various tuned circuits are coupled mechanically so that only one tuning control and dial are required. This implies that whatever the desired signal frequency within the AM broadcast band, the antenna tuned circuit at the input to RF amplifier (if one is provided) or mixer circuit must be tuned to this frequency and the LO must simultaneously be tuned to a frequency which is precisely higher than the signal frequency by the IF frequency of the receiver. Any errors that exist in this frequency difference will result in an incorrect frequency being fed to the IF amplifier and must naturally be avoided. Such errors as exist are called *tracking errors* and they result in stations appearing on either side of their designated positions on the tuning dial.

Maintaining a constant frequency difference between the local oscillator and the RF tuned circuit over the entire broadcast band is possible neither in theory nor in practice and hence some tracking error will always be present. However, with the aid of trimmer and padder capacitors, excellent tracking can be achieved,as shown in Figure 3.13 The trimmer is a small variable capacitor connected in parallel with the variable-tuning capacitor's oscillator section. The trimmer is adjusted to modify the tuning curve of the oscillator to obtain best tracking in the high-frequency portion of the range. The padder is a variable capacitor of relatively large maximum capacitance, connected in series with one of the leads between the tuning-capacitor section and the coil. It is used to make tracking adjustments in the low-frequency portion of the tuning range. If the coil is slug-tuned, a slug adjustment may take the place of the padder capacitor. Figure 3.14 shows the connections of a padder and trimmer in a tuned circuit.

The reasons for the greater effect of the parallel-connected trimmer capacitor on the high-frequency portion of the tuning range, and the series-connected padder on the low-frequency portion, can be understood by considering these conditions separately as follows:

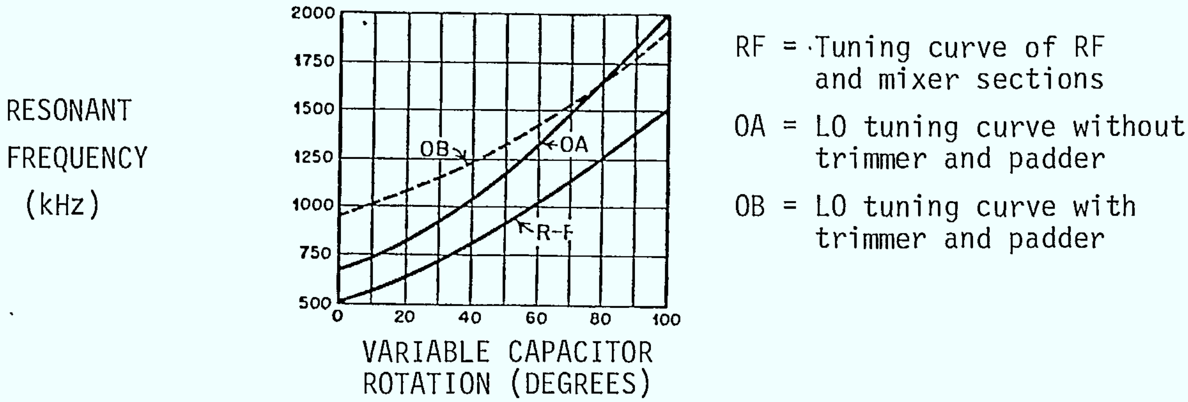


FIGURE 3.13 LOCAL OSCILLATOR TUNING CURVES WITH AND WITHOUT TRIMMER AND PADDING CAPACITANCES

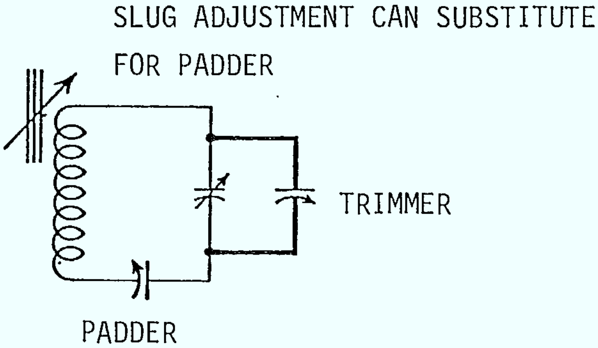


FIGURE 3.14 PLACEMENT OF THE TRIMMER AND PADDER IN A TUNED CIRCUIT

1. <u>High-frequency End of Tuning Range.</u>

For this condition, the main tuning-capacitor section is set for minimum capacitance, which for standard AM broadcast receivers is usually in the order of 5 or 10 pF. The trimmer, parallel-connected and with a maximum capacitance of about 25 pF adds to the capacitance of the tuning-capacitor section to make a total of about 30 or 35 pF. The trimmer thus constitutes most of the capacitance of the circuit, and its adjustment has a very important effect on the resonance frequency.

On the other hand, the padder, which is series-connected with the main tuning-capacitor section, tends to reduce the total capacitance by its addition in the circuit. It is an importance characteristic of two series-connected capacitors of different capacitances that the smaller capacitor exerts a much greater effect on the total capacitance than the larger capacitor. Since the padder ordinarily has a comparatively large value (from 500 to 1000 pF), its effect on the total capacitance of the circuit when the tuning capacitor is set at a low value is negligible.

2. <u>Low-frequency End of Tuning Range</u>

Now the capacitance of the main tuning-capacitor section is set near maximum value, and its value, compared to that of the 25-pF trimmer, is so great that the effect of the trimmer is negligible.

However, the value of the padder relative to the tuning-capacitor section has now become much smaller than at the high-frequency end. It therefore has a more important effect on the resonance frequency now than it has at the high-frequency end of the tuning range.

With AM broadcast receivers, reasonably good reception is possible without a high degree of tracking accuracy. Many receivers of this type include only trimmers and either no padder adjustment at all or a slotted tuning-capacitor plate.

It should be noted here that, fundamentally, trimmer adjustment (high frequency) depends upon capacitance, but padder adjustment effectively adjusts the *inductance* in the circuit. That is why an adjustable iron-dust slug in the coil may be substituted for the padder capacitor.

This fact is important in receivers which do not have padder adjustments, because in these receivers, if the *inductance* is not at the proper value, no amount of trimmer adjustment can make the receiver track. It should be remembered that coil tolerances are sometimes quite large and an inductance error may make tracking adjustments ineffectual.

In some receivers, tracking is achieved mechanically by using the "shaped-plate" method. The local oscillator-tuning capacitor plates are shaped so that the capacitance variation will be proper for tracking as the capacitor shaft is rotated.

3.5.5 Conversion Compression and Intercept Point

In Sections 3.4.4 and 3.4.5, the concepts of the 1 dB gain compression point and third-order intercept point were introduced as a means of describing the large-signal handling capability and intermodulation performance respectively of an amplifier. Increasingly, these concepts are also being used to describe the performance of passive diode mixers, commonly used at VHF and UHF frequencies.

In the case of mixers, it is perhaps more meaningful, however, to speak of a 1 dB *conversion* compression point; this is defined as the point at which the RF input level to the mixer causes the IF output level to deviate from a linear relationship (typically) by 1 dB. In contrast to passive diode mixers for which a conversion *loss* is sustained, the mixers found in AM receivers are active and typically possess a conversion *gain* of between 20 to 30 dB. Since the conversion compression wich occurs indicates saturation of the mixer by the RF input signal, it is clear that a high value for the 1 dB compression point is desirable if saturation, and hence distortion of the recovered baseband audio signal, is to be avoided. This is especially true for the self-oscillating mixer circuit to which AGC cannot be applied and which is seldom preceded by an AGC RF amplifier stage. A (two-tone) third-order intercept point may also be determined for mixers. As in the case of an amplifier, the intercept point may be used to predict the level of second and third order IM products present in the IF output of a mixer in its linear region of operation (i.e., below the 1 dB compression point). Since harmonic distortion in the audio output is caused primarily by the generation of intermodulation products in the mixer, a high value for the 3rd-order intercept point and operation well below the 1 dB compression point (i.e., in the linear region of operation) are thus desirable in minimizing distortion due to this source.

It is perhaps worthy to note here that neither of the above concepts are presently used in describing the performance of AM broadcast receivers. Rather, performance is usually defined on the basis of the input RF level required to produce a harmonic distortion of 10% in the audio output of the receiver (overload response) at a modulation depth of 30%.

3.5.6 Spurious Responses

Spurious external frequencies reaching the RF input of the mixer can result in the generation of undesired frequencies that may fall in the passband of the IF stage. The interference manifests itself in the audio output as a whistle or a "tweet" which consists of an audio note whose pitch varies in a sirenlike manner, passing from a high audio frequency down through "zero beat" and back up again as the tuning dial of the receiver is rotated. All whistles result from the existence of two different IF's at the input to the envelope detector, the two frequencies having a difference which varies with the change of the local oscillator frequency.

The condition for interference in the IF passband is:

$$mf_i \pm nf_{LO} = f_{IF} \tag{3.7}$$

where

f_i = spurious frequencies at the RF input to the mixer

f_{LO} = local oscillator frequency

f_{IF} = intermediate frequency of the receiver

m, n = integers

Figure 3.15 is a chart for identifying whistles heard by a receiver whose IF frequency is 455 kHz.

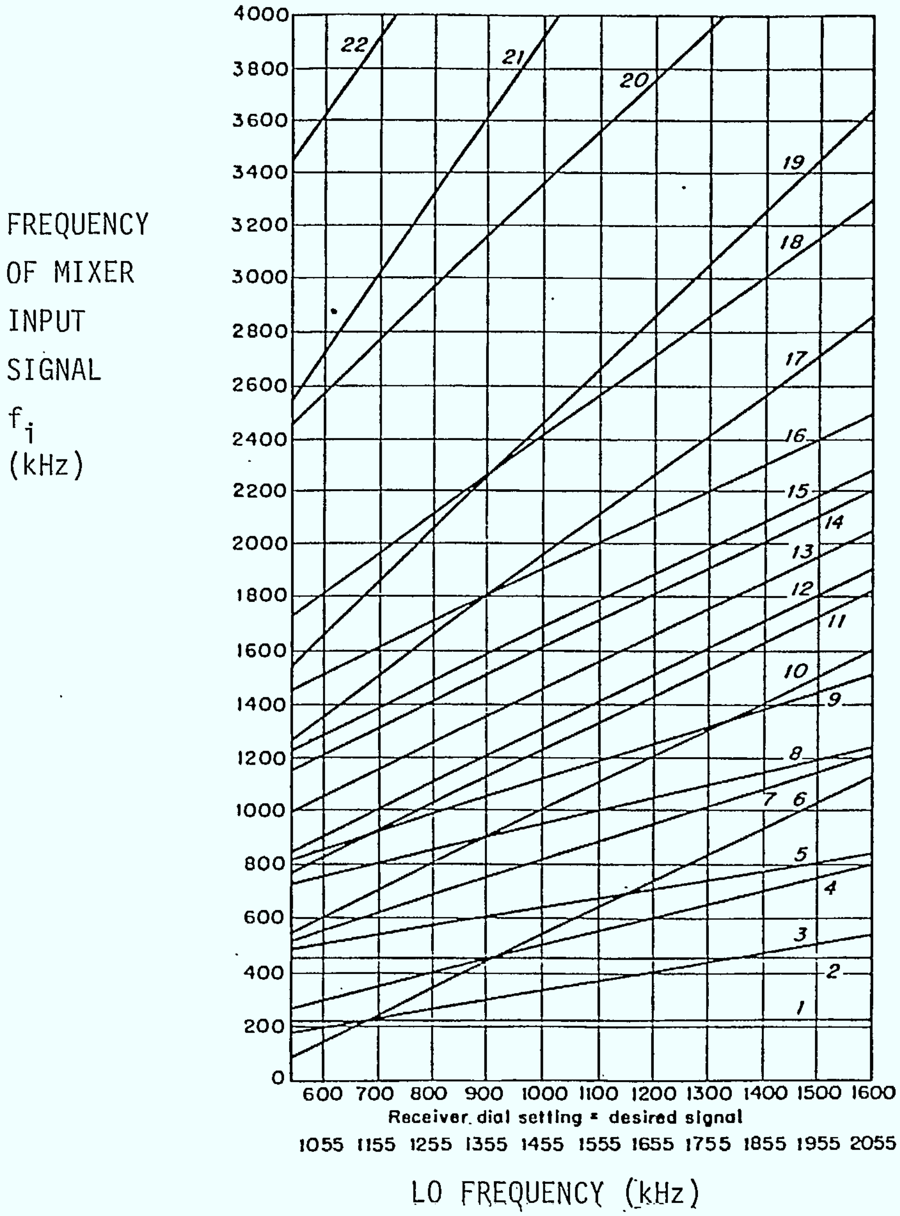
Whistles are worst under the following conditions:

1. Limited selectivity ahead of the mixer;
2. Small values of m, n involved;
3. Strong interfering signals.

Curve 13 is included in Figure 3.15 for interference equal to the local oscillator frequency. Such interference amounts to another oscillator and another IF whose frequencies do not change as the receiver dial is slightly rotated. The regular IF will, however, change with dial rotation hence producing a whistle. Curve 6 is for interference located 455 kHz below the signal to which the receiver is tuned and is only of importance in the case of a strong interfering signal. Curves 2 and 1 illustrate the susceptibility of a receiver to whistles resulting from direct IF interference and at half the IF respectively. Note that the LO is not involved in these two cases.

Consider, as a hypothetical example, a case in which there is a strong local station in the broadcast band at 810 kHz and a weak distant station at 580 kHz. A receiver is tuned to the distant station and a whistle at 5 kHz is heard on the receiver. If we locate f_i = 810 kHz on the ordinate and the desired signal on the abscissa at 580 kHz of Figure 3.15 , we find that this whistle corresponds closest to the straight line graph labelled (11) for which

$$2\, f_{LO} - 2f_i = 455 \text{ kHz}$$



(22)	$f_i - 3f_{LO}$	= 455 kHz
(21)	$3f_{LO} - f_i$	= 455 kHz
(20)	$f_i - 2f_{LO}$	= 455 kHz
(19)	$2f_{LO} - f_i$	= 455 kHz
(18)	$2f_i - 3f_{LO}$	= 455 kHz
(17)	$3f_{LO} - 2f_i$	= 455 kHz
(16)	$f_i - f_{LO}$ (image)	= 455 kHz
(15)	$2f_i - 2f_{LO}$	= 455 kHz
(14)	$3f_i - 3f_{LO}$	= 455 kHz
(13)	$f_{LO} - f_i$	= 0
(12)	$3f_{LO} - 3f_i$	= 455 kHz
(11)	$2f_{LO} - 2f_i$	= 455 kHz
(10)	$f_{LO} - f_i$ (desired)	= 455 kHz
(9)	$3f_i - 2f_{LO}$	= 455 kHz
(8)	$2f_i - f_{LO}$	= 455 kHz
(7)	$2f_{LO} - 3f_i$	= 455 kHz
(6)	$f_{LO} - f_i$	= 910 kHz
(5)	$3f_i - f_{LO}$	= 455 kHz
(4)	$f_{LO} - 2f_i$	= 455 kHz
(3)	$f_{LO} - 3f_i$	= 455 kHz
(2)	f_i	= 455 kHz
(1)	$2f_i$ or f_i	= 455 kHz or = 227.5 kHz

FIGURE 3.15 CHART FOR IDENTIFYING WHISTLES HEARD WITH A RECEIVER WHOSE INTERMEDIATE FREQUENCY IS 455 kHz

For a desired signal at 580 kHz, the local oscillator frequency required is:

$$f_{LO} = 580 + 455 = 1035 \text{ kHz}$$

and

$$f_i = 810 \text{ kHz}$$

therefore

$$2f_{LO} - 2f_i = 2070 - 1620 \text{ kHz}$$

$$= 450 \text{ kHz}$$

The interference at 450 kHz then mixes with the 455 kHz desired signal in the baseband envelope detector to produce the 5 kHz whistle.

Whistles may also occur when one is listening to a station near a harmonic of the IF frequency, without the presence of an interfering signal. For a typical IF of 455 kHz, such whistles are likely to interfere with reception at 910 kHz (2 x IF) and 1365 kHz (3 x IF). A whistle can occur at these locations because of the presence of two slightly different IF's when the receiver is slightly detuned, as mentioned earlier.

A recent case in point was the CBC AM radio broadcast in Ottawa which had been allocated the 910 kHz channel but was forced to shift to the adjacent 920 kHz channel because of the annoying whistle interfering with reception.

To understand why the whistle occurs, let us assume that the desired signal is at 910 kHz and the receiver's IF is 455 kHz. If the local oscillator is slightly detuned to 1366 kHz instead of 1365 kHz, the two IF's obtained are:

1. A desired IF = 1366 - 910 kHz = 456 kHz

2. An undesired IF due to the mixing which takes place between the second harmonic of the signal and the local oscillator, that is:

$$
\begin{aligned}
\text{Undesired IF} &= 2 \times 910 - 1366 \text{ kHz} \\
&= 1820 - 1366 \text{ kHz} \\
&= 454 \text{ kHz}
\end{aligned}
$$

As before, these two IF frequencies combine in the envelope detector to produce a 2 kHz whistle. To reduce the annoyance of the whistle somewhat, the simplest solution would be to slightly shift the IF frequency. A more satisfactory but more costly solution would be to use a balanced mixer to suppress the second harmonic of the signal. In some AM receiver integrated circuits (IC's), a balanced mixer is incorporated to suppress whistles that occur at signal frequencies that are harmonics of the IF frequency.

Another possible source of whistle at 910 kHz is a feedback of harmonics of the IF from the envelope detector into the RF circuitry. This source of interference is indicated when a measurable amount of whistle is still present even under low signal input conditions. The remedy is to rearrange the wiring or provide shielding or bypassing to remove the feedback. In some cases, feedback may occur through the AGC bias lead to the RF or mixer stages.

3.5.7 Local Oscillator Radiation and Drift

Spurious Radiation

LO radiation from an AM receiver may be experienced either as a result of direct radiation from the LO circuitry itself or by radiation from the receiver's antenna as a result of poor isolation between the LO and antenna. LO interference with other broadcast receivers may occur in a situation where two stations in the same area are separated by the IF frequency of 455 kHz (refer to Section 3.4.6, Figure 3.15 , curve 6). To prevent such a situation arising, the frequencies of broadcasting stations in the same area are separated by intervals either less than 450 kHz or greater than 460 kHz. However, with the present congestion in the AM band and siting difficulties in certain areas, this may not always be possible to achieve.

On such occasions where there is a possibility of LO interference due to the overlapping of the stations' 0.5 mV/m field strength contours, the DOC requires that any complaints of LO interference to the reception of an existing station be investigated and rectified (refer to Broadcast Procedure 1, Rule 11).

In the U.S.A., the Federal Communications Commission (FCC) requires that the field strength of spurious LO radiation at a distance of 100 ft. or more from a receiver be less than 100 μV/m for frequencies in the range from 0.45 MHz up to and including 25 MHz.

LO reradiation from a receivers antenna can be significantly reduced by using either;

(a) an input RF amplifier;

(b) a balanced mixer, or

(c) a dual insulated-gate MOSFET transistor (provides high isolation between the LO and RF input ports) as a mixer.

In the case of direct radiation from the LO circuitry, proper shielding will be required to minimize this mode of spurious radiation.

Drift

Any drift in the frequency of the local oscillator due to temperature variations will result in an IF of the improper frequency emerging from the mixer. If severe, loss of signal reception will occur due to the attenuation that will be sustained in the IF section as the signal drifts outside the IF passband. LO drift will thus require the listener to frequently readjust his tuning dial to maintain proper reception which could be extremely annoying.

LO frequency drift in AM receivers is seldom encountered in practice since the tuning frequencies are largely determined by passive machanical components and not by active circuitry capacitances, which are usually very temperature and voltage sensitive. If a highly stable LO source is necessary, which it rarely is, one might resort to use of frequency locking techniques (e.g., frequency synthesizer) which are more complicated and costly. Such approaches might only be contemplated for the premium-priced receivers (e.g., console-type AM/FM stereo receivers).

With respect to LO drift due to battery voltage, there were no problems of serious concern that could readily be identified.

3.6 IF Stage

Numerous factors must be considered in the design of the IF amplifier stages such as the choice of IF frequency, the number of stages, gain and selectivity required.

3.6.1 Choice of IF Frequency

The chief considerations in the choice of an IF frequency are:

1. If the IF frequency is too high, poor selectivity and poor adjacent-channel rejection result;

2. A high value of IF frequency increases tracking difficulties;

3. As the IF frequency is lowered, image-frequency rejection becomes poorer. Equation 3.6 showed that as the ratio of image frequency to signal frequency is increased, image rejection improves and this, naturally, favours a high IF frequency;

4. A very low IF frequency makes the selectivity too sharp which results in attenuation of the upper sidebands of a signal. This problem arises because the Q has to be low when the IF is low, and

hence the gain per stage is low; thus a designer is more likely to raise the Q than to increase the number of IF amplifiers;

5. If the IF is very low, the frequency stability of the local oscillator must be made correspondingly higher because any frequency drift is now a larger proportion of a low IF than of a high IF;

6. The IF frequency must not fall within the tuning range of the receiver or else instability will occur and whistles will be heard, making it impossible to tune to the frequency band immediately adjacent to the IF frequency.

For AM receivers, the most common IF frequency encountered is 455 kHz although a variety of different IF frequencies in the range 438 to 465 kHz may also be found. For automotive AM receivers, either 455 kHz or 262.5 kHz may be used.

For an IF frequency of 455 kHz, the image frequency will occur at 1450 kHz for a receiver tuned to 540 kHz, thus allowing good image rejection with a simple RF tuned circuit. Since image channel and direct IF rejection is performed by the RF tuned circuits, a receiver which uses an IF frequency different from 455 kHz but in the range 438 to 465 kHz will still experience adequate image channel rejection because of the nearly 1 MHz separation between the desired channel and its image.

A receiver with a higher IF frequency will, however, tend to be more susceptible to direct IF interference since it becomes increasingly difficult for a simple RF tuned circuit to provide adequate rejection when tuned to the lower frequency end of the AM band, e.g., 540 kHz. It thus seems that there is no strong technical ground for dictating the use of any one particular IF frequency since the choice of an IF frequency in the range 438 to 465 kHz imposes no restrictions on the way in which image channels can be assigned. Rather, the increasing trend towards the use of a common 455 kHz IF will be dictated by economic considerations (e.g., mass production of 455 kHz IF circuit elements).

There could be some merit, however, in considering the use of a 450 kHz IF instead since this particular IF would ease the implementation of local oscillators using frequency synthesizers without significantly affecting current AM receiver designs. The reason is that, with all synthesis techniques, locking of the LO to multiples of the RF channel spacing implies that the IF frequency is also at a multiple of the same channel spacing. For AM receivers with an IF frequency of 450 kHz, both a 10 kHz and a narrower 9 kHz spacing could be used. Since the synthesizer reference frequency is usually derived by division of a crystal oscillator, a change of channel spacing from 10 kHz to 9 kHz would simply entail the changing of a crystal.

If, however, the IF is constrained to be 455 kHz, then for 10 kHz channel spacings, the reference frequency must be

5 kHz. The synthesizer would then be allowed to lock on every other multiple. With a 9 kHz spacing, a reference frequency of 1 kHz would be required. Neither reference frequency is desirable since the inevitable frequency synthesizer sidebands at 5 kHz or 1 kHz will tend to cause spurious audio whistles.

3.6.2 Stage Gain

The IF amplifier stage is primarily responsible for the gain and selectivity (adjacent channel rejection) of the receiver. Since the IF stage operates at a fixed frequency, the bandwidth of the receiver is determined by the IF stage and is independent of receiver tuning over the entire range from 540 to 1600 kHz.

Although it is tempting to design essentially all of the receivers gain into the IF stage, this is not desirable in practice since instability may result from feedback unless expensive shielding is provided. Consequently, the receiver's total gain is distributed between the RF, IF and audio amplifier stages. A typical gain distribution is as follows: 10 to 30 dB in the RF stage, 60 dB in the IF stage and 30 to 60 dB in the audio section.

IF designs usually employ two stages of single-tuned amplification to obtain a gain of 60 dB. Figure 3.16 illustrates a typical 2-stage transistor IF amplifier with AGC applied to the first stage.

Although a double-tuned circuit rejects adjacent frequencies far better than a single-tuned circuit, transistor amplifiers appear, on the whole, to use

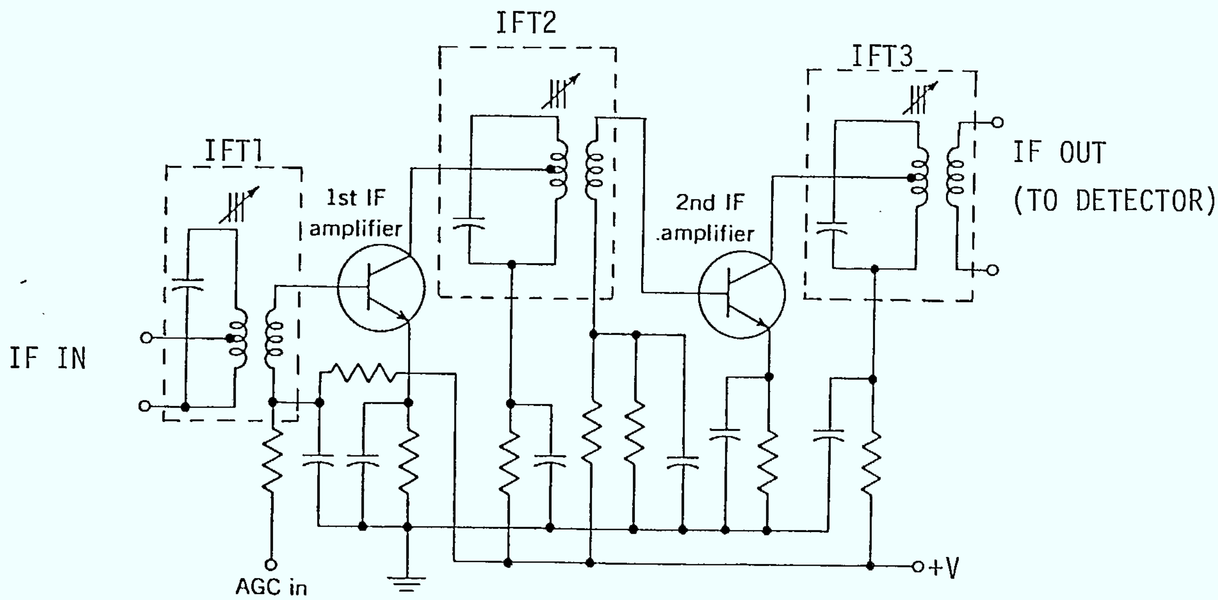


FIGURE 3.16 TRANSISTOR TWO-STAGE IF AMPLIFIER

single-tuned circuits for interstage coupling. The reason is simply that greater gain may be achieved in this way because of the need for tapping coils in tuned circuits. This tapping may be required to obtain maximum power transfer and a reduction of the damping of the circuit involved. It will be recalled that the bandwidth of a tuned circuit depends on its loaded Q, which depends on the unloaded Q and the external damping resistance. Since transistor impedances may be low, tapping is used. If a double-tuned transformer is used, both sides of it might have to be tapped, rather than just one side as with a single-tuned transformer; there would thus be a reduction in voltage applied to each transistor and hence a general reduction in gain. To facilitate interchangeability, IF transformers are often made all identical.

3.6.3 Selectivity And Bandwidth

The adjacent channel selectivity of an AM broadcast receiver is the extent to which it is able to reject an unwanted carrier (interference) which may be located in a channel adjacent to the one to which the receiver is tuned. It is expressed either in terms of decibels or as the corresponding voltage ratio. Clearly the adjacent channel selectivity of an AM receiver is a key parameter in the efficient utilization of the AM band since the greater the selectivity or discrimination, the closer channels may be assigned within a given area without causing mutual interference.

The selectivity at 10 kHz away from the channel to which a receiver is tuned is usually referred to as the adjacent-channel attenuation (ACA); at 20 kHz away, it is referred as the second channel attenuation (SCA). Typical values for the ACA and SCA are 25 dB and 52 dB respectively.

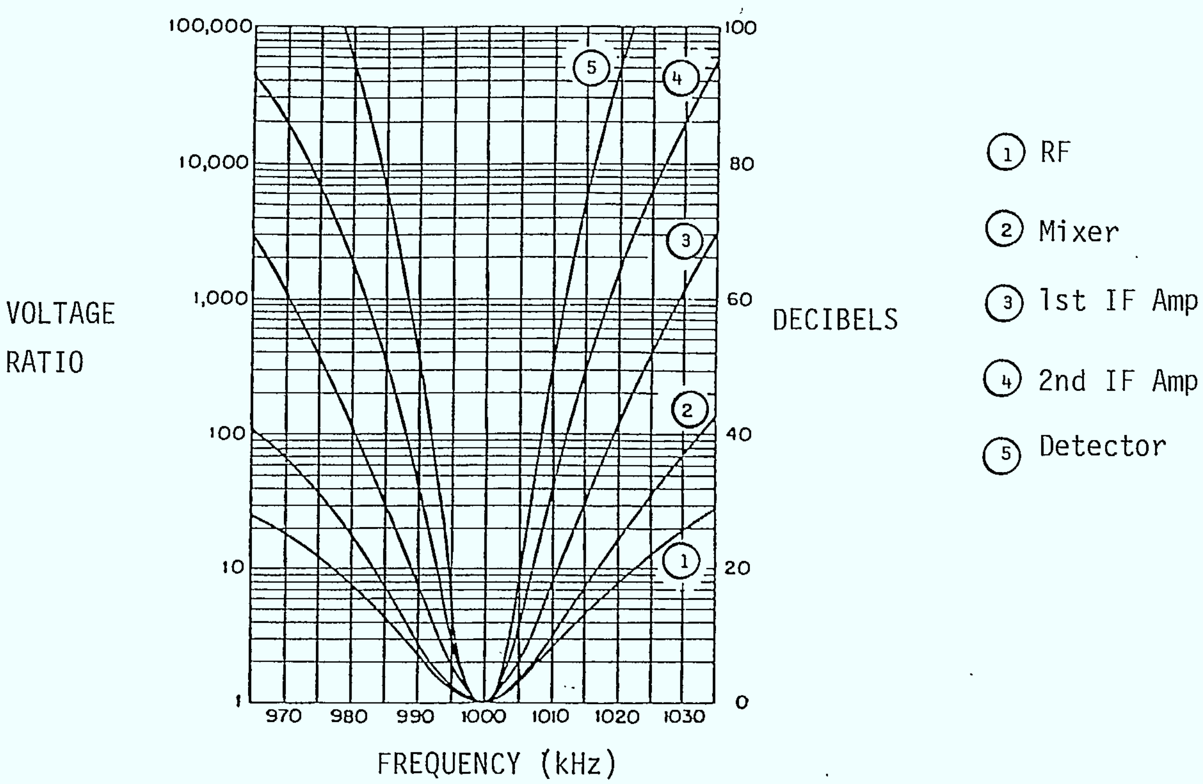


FIGURE 3.17 INCREASING SELECTIVITY IN AN AM RECEIVER

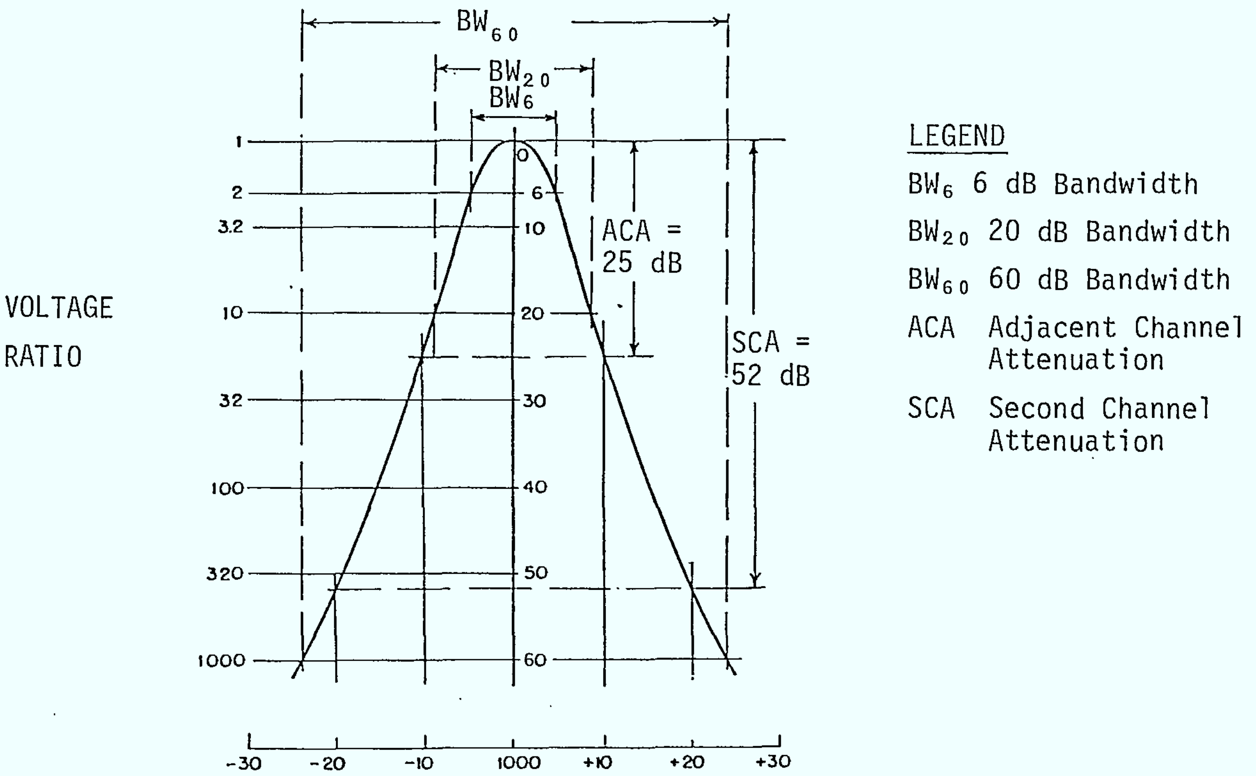


FIGURE 3.18 TYPICAL SELECTIVITY CURVE FOR AN AM RECEIVER

Figure 3.17 illustrates the increasing selectivity which might typically be obtained at various stages of an AM receiver equipped with two stages of IF amplification while Figure 3.18 illustrates the typical selectivity curve of a complete AM receiver.

While it is desirable to provide as much selectivity as possible, it is also important to ensure that the bandwidth or passband of the receiver is not unduly reduced in the process. The reason is that by reducing the passband, the AM carrier sidebands will be attenuated, resulting in the loss of the higher frequency components in the demodulated signal. For speech programs, reducing a receiver's bandwidth form 10 kHz to 6 kHz would not greatly affect received audio signal quality since speech contains few frequency components above 3 kHz. On the other hand, musical programs would suffer greatly since they can contain frequency components as high as 15 kHz.

Note in Figure 3.18 that the 6 dB bandwidth (BW_6) or passband of the receiver is about 9 kHz. In filter terminology, the ratio of the 60 dB to the 6 dB bandwidth is often referred to as the *shape factor*; hence, the smaller the shape factor, the greater the adjacent channel attenuation or selectivity of the receiver.

3.6.4 AGC

Saturated operation of the IF stage is prevented by applying a negative AGC voltage derived in the envelope detector to the first IF amplifier stage (Figure 3.16) Significant gain reduction occurs when the collector current is reduced and the base and emitter voltages approach zero (i.e., the transistor stage approaches cutoff). An AGC gain control of over 40 dB can be obtained with the single stage. Typical AGC begins to operate at about 1 mV from the detector and continues to reduce the IF gain to maintain a relatively constant IF signal level at the input to the envelope detector.

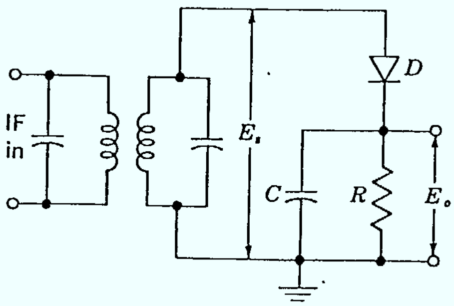
3.7 Detector Stage

3.7.1 Envelope Detection

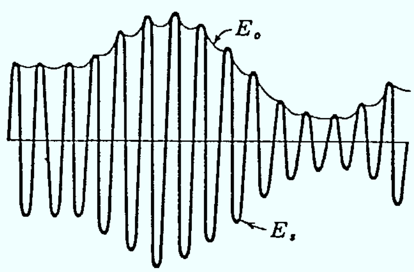
For envelope detection of AM broadcast signals, the series-connected diode detector is almost universally employed because of its simplicity, efficiency and relatively low distortion when properly designed.

Figure 3.19 (a) illustrates the basic form of the diode detector circuit. On positive-going excursions of the carrier wave, the diode conducts and charges capacitor C to the peak value of the carrier. On negative-going excursions, the diode is reverse-biased (i.e., in cut-off) and the capacitor begins to discharge through resistor R. On the next positive-going cycle the diode again conducts and restores the charge on C, as shown in Figure 3.19 (b).

Thus, for an unmodulated or pure-carrier input, the voltage across C is restored to peak value once per carrier cycle. If the time constant RC is long compared with the period of the carrier wave, there is little discharge of C during negative-going half-cycles of the carrier and the voltage across C then approximates the peak value of the carrier signal input. Note that the diode detector's output has an average or d.c. value which is proportional to the carrier level and this fact is used in some receivers for automatic gain control or AGC (discussed later in this section). Because of the slight discharge of the capacitor during the negative-going half-cycles of the carrier wave, there is also a residual RF ripple present whose frequency is twice the IF frequency or, typically, 910 kHz. The d.c. component is



(a) Circuit Diagram



(b) Input And Output Voltages

FIGURE 3.19 SIMPLE DIODE DETECTOR
(a) CIRCUIT DIAGRAM
(b) INPUT AND OUTPUT VOLTAGES

prevented from reaching the audio amplifier stage following the diode detector by capacitively coupling the two stages while the RF signal is removed by a low-pass filter.

A more practical version of the diode detector is shown in Figure 3.20 . The resistor R of Figure 3.19 (a) has been split into two resistors (R_1 and R_2) to ensure that there is a series d.c. path to ground for the diode while a low-pass filter (comprised of R_1 - C_2) has been added to remove the residual RF ripple in the diode output. C_3 is a coupling or blocking capacitor whose function is to prevent the diode's d.c. output from reaching the audio volume control R_3. The low pass (AGC) filter comprised of R_4 - C_4 is designed to remove the audio frequencies from the diode output thus providing a d.c. voltage (positive, in this case) for automatic gain control (AGC). In cases where a negative AGC voltage is desired instead, the diode in Figure 3.20 is simply reversed.

From Figure 3.20 , it is clear that the d.c. diode load, R_c, is equal to $R_1 + R_2$ whereas the a.c. load impedance, Z_m, is equal to R_1 in series with the parallel combination of R_2, R_3 and R_4, assuming the capacitors have reactances which may be ignored. This assumption is strictly true only at medium frequencies; at high audio frequencies, Z_m may acquire a reactive component which can cause distortion as well as an uneven frequency response.

The IF transformer shown in Figure 3.20 is used for coupling the IF stages to the diode detector and could instead be replaced by a narrowband bandpass filter, for example.

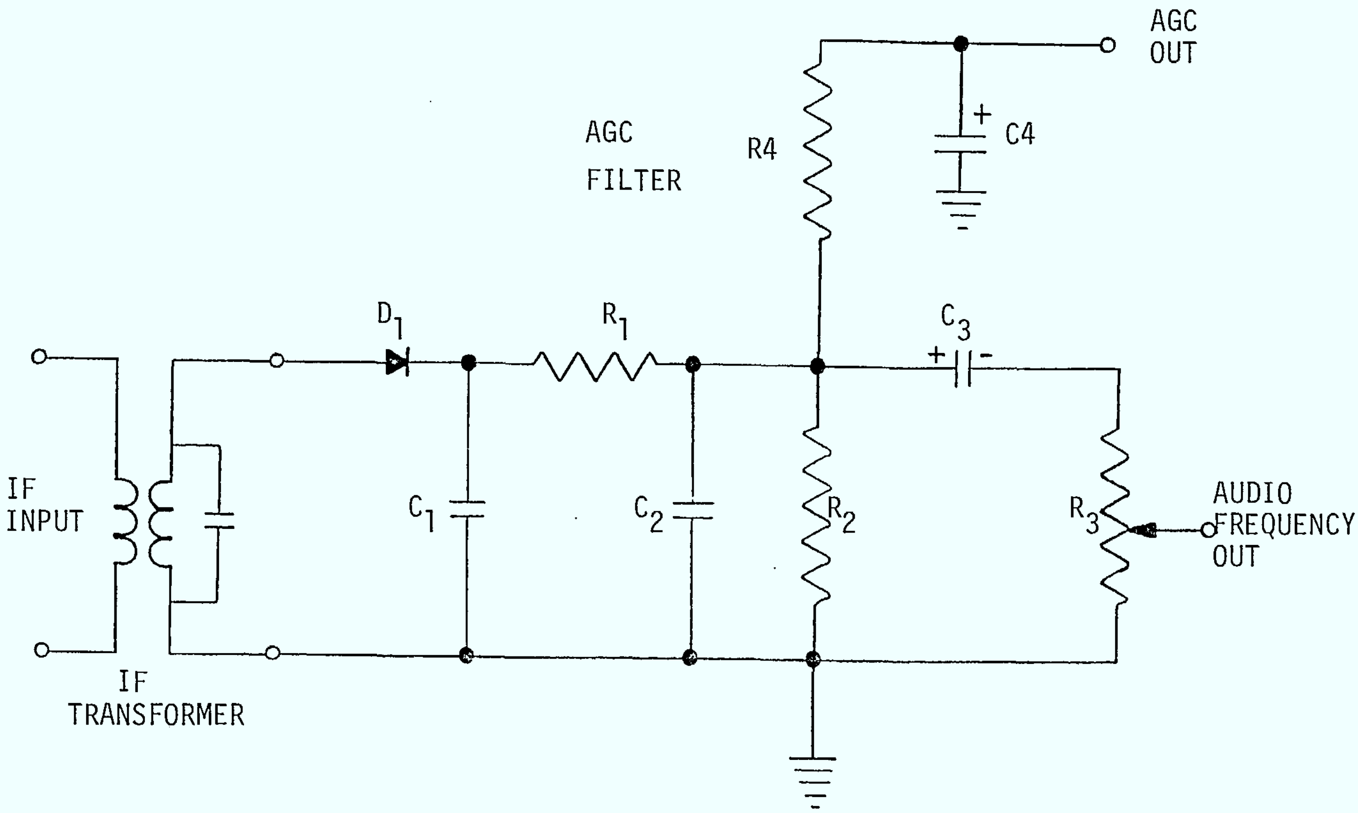


FIGURE 3.20 A PRACTICAL DIODE DETECTOR

Distortion

There are basically three sources of distortion in diode peak detectors: the first is due to the inherent nonlinearity of the diode's response itself, the second is caused by the fact that the a.c. and d.c. diode load impedances are not equal and the third, the fact that the a.c. load impedance acquires a reactive component at the highest audio frequencies.

To demonstrate the distortion caused by the nonlinear response of the diode, let us consider the case of a cosinusoidally modulated AM carrier. The AM carrier signal is then given by (Equation 2.4, Section 2.1):

$$\phi_{AM}(t) = A_c \cos\omega_c t + \frac{mA_c}{2}\cos(\omega_c + \omega_m)t + \frac{mA_c}{2}\cos(\omega_c - \omega_m)t \qquad (3.8)$$

where

A_c = unmodulated carrier amplitude

ω_c = carrier frequency (angular)

ω_m = modulating signal frequency (angular)

m = index of modulation (or modulation depth) $= \frac{A_m}{A_c}$

The current, i(t), developed in a diode detector as a result of an applied voltage, e(t), may be represented as a power series thus:

$$i(t) = a_0 + a_1e(t) = a_2e^2(t) + a_3e^3(t) + \ldots.. \tag{3.9}$$

The only term of interest for the process of detection is the term $a_2e^2(t)$. Substituting $\phi_{AM}(t)$ for $e(t)$ in this term gives

$$\begin{aligned} a_2e^2(t) &= a_2[A_c\cos\omega_c t + \frac{mA_c}{2}\cos(\omega_c+\omega_m)t + \frac{mA_c}{2}\cos(\omega_c-\omega_m)t]^2 \\ &= a_2A_c^2\cos^2\omega_c t + a_2\frac{m^2A_c{}^2}{4}\cos^2(\omega_c+\omega_m)t \\ &\quad + a_2\frac{m^2A_c{}^2}{4}\cos^2(\omega_c-\omega_m)t \\ &\quad + a_2mA_c^2\cos(\omega_c+\omega_m)t\cos\omega_c t \\ &\quad + a_2mA_c^2\cos(\omega_c-\omega_m)t\cos\omega_c t \\ &\quad + a_2\frac{m^2A_c^2}{2}\cos(\omega_c+\omega_m)t\cos(\omega_c-\omega_m)t \end{aligned} \tag{3.10}$$

The first three terms of equation 3.10 produce the frequencies $2f_c$, $2f_c + 2f_m$ and $2f_c - 2f_m$ which are all rejected by the lowpass filter following the diode. The fourth term yields a frequency sum component of $2f_c + f_m$, which is rejected by the filter, and a difference component of $\frac{1}{2}\, a_2mA_c^2\cos\omega_m t$.

The sum component of the fifth term is $2f_c - f_m$, which is rejected by the filter, and the difference component is:

$$\frac{1}{2}a_2 m A_c^2 \cos(-\omega_m)t = \frac{1}{2}a_2 m A_c^2 \cos\omega_m t$$

The sum component of the last term is $2f_c$, which is also rejected, and the difference term is

$$\frac{1}{4}a_2 m^2 A_c^2 \cos 2\omega_m t.$$

Thus the output of the lowpass filter is given by the sum of the unrejected terms, i.e.:

$$e_{out} = a_2 m A_c^2 \cos\omega_m t + \frac{1}{4} a_2 m^2 A_c^2 \cos 2\omega_m t \qquad (3.11)$$

The first term is the output signal. It is properly proportional to m and is of the proper frequency f_m. The second term contains the second harmonic of the modulating signal frequency and represents the harmonic distortion inherent in square-law detection. When the modulation depth is 100%, the harmonic distortion present is 25%.

Just as the modulation index of the modulated carrier signal was defined as the ratio A_m/A_c, we can also define a modulation index of the demodulated wave as:

$$m_d = \frac{I_m}{I_c} \tag{3.12}$$

Note that the definition here is in terms of current since the diode is a current device. Bearing in mind that these currents are peak rather than r.m.s. values, we have:

$$I_m = \frac{A_m}{Z_m} \tag{3.13a}$$

and $$I_c = \frac{A_c}{R_c} \tag{3.13b}$$

where Z_m = diode load (audio) impedance (assumed resistive)

R_c = diode d.c. load resistance

Since the a.c. load impedance is smaller than the d.c. resistance, the audio frequency current, I_m, will be larger than the d.c. current, I_c. Or, alternatively, <u>the modulation index of the demodulated signal is higher than it was for the modulated carrier signal applied to the</u> detector. This clearly suggests that it is possible for overmodulation to occur in the output of the detector although the modulation index of the applied voltage is less than 100%. When the input modulation index is too high for the diode to handle, negative peak clipping of the diode output occurs.

The maximum value of applied modulation index which a diode detector will handle without negative peak clipping is determined in the following manner. The modulation index of the demodulated signal is given by

$$m_d = \frac{I_m}{I_c} = \frac{A_m/Z_m}{A_c/R_c} = \frac{mR_c}{Z_m} \qquad (3.14)$$

Since the maximum tolerable modulation index in the diode output is unity, the permissible modulation index for the transmitted carrier signal is bounded by

$$m < m_{d,max.} = \frac{Z_m}{R_c}$$

i.e.,

$$m < \frac{Z_m}{R_c} \qquad (3.15)$$

At the higher modulating frequencies, Z_m can no longer be assumed to be purely resistive since it acquires a reactive component due to capacitors C, and C_2 (refer to Figure 3.20). At high modulation depths, the diode output current will be changing so quickly that the discharge time constant of the load may be unable to follow the change. As a result, the diode current will instead decay exponentially, as illustrated in Figure 3.21. This type of distortion is referred to as *diagonal clipping*. Since it does not normally occur when the modulation depth, m_d, is below approximately 60%(at the highest modulating frequency), it is possible to design a diode detector without this type of distortion.

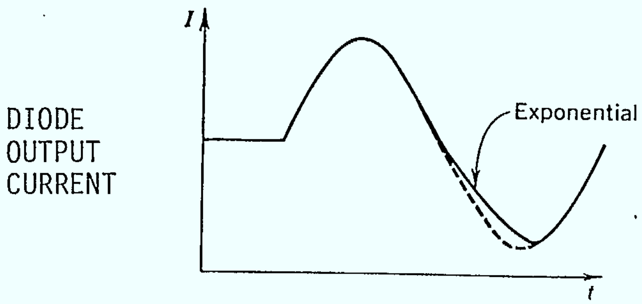


FIGURE 3.21 DIAGONAL CLIPPING

3.7.2 Rectifier Detection

Figure 3.22 illustrates a simplified block diagram of a rectifier detector with the waveforms present at various stages of the detector. The diode rectifies the IF input signal by removing the negative half-cycles of the carrier wave. This is equivalent to multiplying the input signal by unity for positive-going half-cycles and zero for negative-going half-cycles of the carrier wave. Rectification can thus be viewed as being equivalent to multiplying the modulated carrier signal by a square wave p(t) of frequency ω_c. This being the case, the spectrum of the rectified signal is therefore obtained by convolving the spectrum of the modulated signal (Figure 3. 23 b) with that of p(t) (Figure 3. 23 c).

The result of the graphical convolution is shown in Figure 3.23f , and it is evident that the baseband signal f(t) can be recovered by lowpass filtering the output of the rectifier. The output of the lowpass filter, however, still contains a d.c. term (impulse at the orgin - see Figure 3.23f) which can be removed from the baseband output by a blocking capacitor.

Thus, rectifier detection is in essence a synchronous form of detection since the operation of rectification is equivalent to multiplication of the modulated carrier signal by a periodic signal (a square wave of frequency ω_c). Note, however, that multiplication is performed without any carrier signal and is a result of the high carrier content in the modulated signal itself. If there were no carrier present (as in the case of a suppressed carrier), then the rectification operation would no longer be equivalent to multiplication of

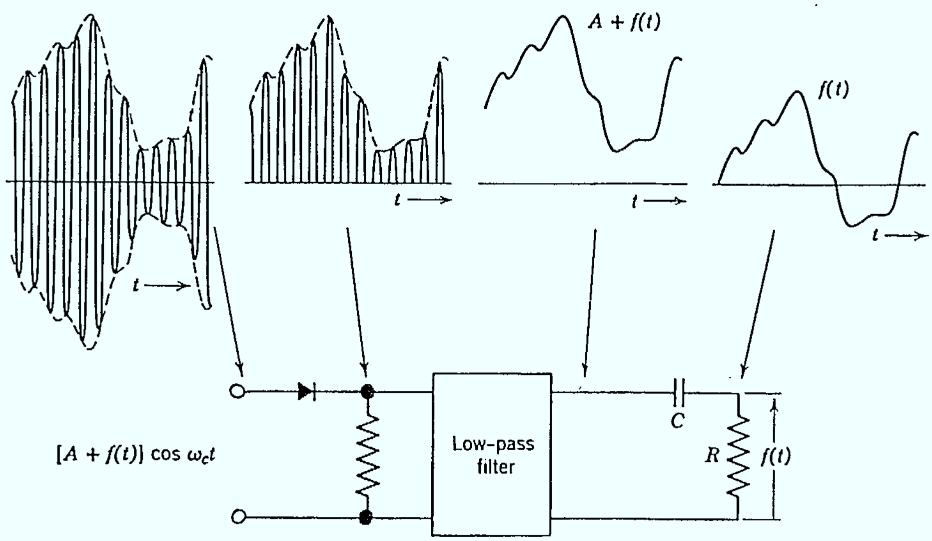


FIGURE 3.22
SIMPLIFIED BLOCK DIAGRAM
OF A RECTIFIER DETECTOR

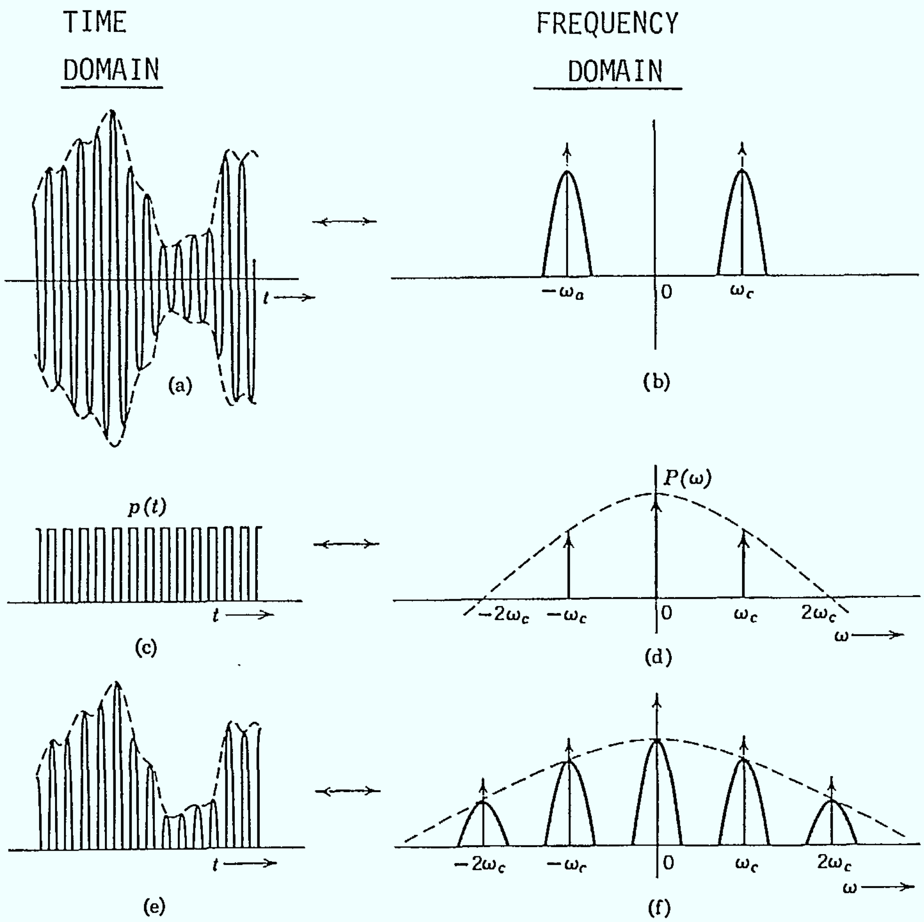


FIGURE 3.23

TIME AND FREQUENCY DOMAIN

SIGNAL REPRESENTATIONS FOR

THE RECTIFIER DETECTOR

the modulated signal by the square wave, p(t). As in the case of the envelope detector, it must ensured that the unmodulated carrier amplitude, A_c, is greater than or equal to the minimum absolute value of f(t), i.e., $A_c > |\min. f(t)|$, in order for rectifier detection to occur. If overmodulation (i.e., $A_c < |\min. f(t)|$ were to occur, severe baseband distortion would result since the rectifier detector would be unable to recover f(t) for the reason given above in the case of suppressed carrier transmission.

3.7.3 Synchronous Detection

In order to synchronously or coherently demodulate an AM signal, the incoming modulated carrier signal is mixed with a local oscillator which is exactly coherent with the incoming carrier (see Figure 3.24). Any frequency errors between the local oscillator and the received carrier will result in serious distortion of the recovered baseband signal, as shown below.

As before, the AM signal, $\phi_{AM}(t)$, is given by

$$\phi_{AM}(t) = [A_c + f(t)] \cos \omega_c t \tag{3.16}$$

Let us assume that, in the general case, the local oscillator is $\cos(\omega_c + \Delta\omega)t$, where $\Delta\omega$ represents a small frequency error. The output of the mixer or product detector is therefore:

$$\begin{aligned}\phi_{AM}(t)\cos(\omega_c+\Delta\omega)t &= [A_c+f(t)]\cos \omega_c t \times \cos(\omega_c+\Delta\omega)t \\ &= A_c \cos \omega_c t \cos(\omega_c + \Delta\omega)t \; + \\ &\quad f(t)\cos \omega_c t \cos(\omega_c + \Delta\omega)t\end{aligned} \tag{3.17}$$

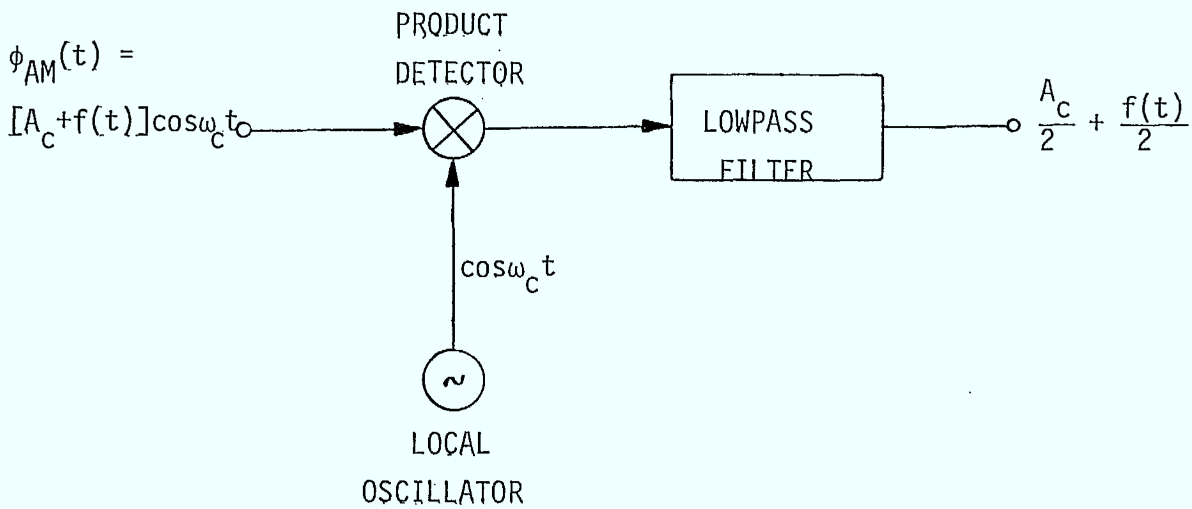


FIGURE 3.24 SYNCHRONOUS DETECTION OF AN AM SIGNAL

By resorting to trigonometric identities, the right-hand side of equation (3.17) may be expressed as:

$$\text{R.H.S.} = \frac{A_c}{2}[\cos(2\omega_c - \Delta\omega)t + \cos\Delta\omega t] + \frac{f(t)}{2}[\cos(2\omega_c + \Delta\omega)t + \cos\Delta\omega t] \tag{3.18}$$

The lowpass filter at the output of the product detector removes those terms containing $\cos(2\omega_c + \Delta\omega)t$ and we are left with:

$$\text{Lowpass Filter (LPF) Output} = \frac{A_c}{2}\cos\Delta\omega t + \frac{f(t)}{2}\cos\Delta\omega t \tag{3.19}$$

It is clear from equation (3.19) that instead of recovering the baseband signal $f(t)$, we obtain a distorted version given by $\frac{f(t)}{2}\cos\Delta\omega t$ $(\Delta\omega \neq o)$. If we can ensure that $\Delta\omega = o$ then equation (3.19) reduces to the desired ideal form:

$$\text{LPF Output} = \frac{A_c}{2} + \frac{f(t)}{2} \tag{3.20}$$

As in the two previous detection techniques, the d.c. term ($\frac{A_c}{2}$, in this case) is removed from the baseband output to the audio amplifier stage by a blocking capacitor. An AGC control voltage may similarly be derived from the LPF output by filtering out the signal component $\frac{f(t)}{2}$ instead in order to obtain the d.c. component, $\frac{A_c}{2}$.

Ensuring that $\Delta\omega$ is always equal to zero requires fairly elaborate and expensive circuitry which is the reason why synchronous detection is rarely, if ever, implemented in even the most expensive AM broadcast receivers.

3.7.4 Automatic Gain Control

Automatic gain control (AGC) is a method by which the overall gain of a receiver is automatically varied with the changing strength of the received carrier signal to maintain a relatively constant output (d.c.) level from the detector. AGC also averts nonlinear operation of the IF amplifier(s) by ensuring that the amplifiers are not driven into saturation by abnormally high carrier levels. Saturated operation can lead to the unwanted generation of harmonics and intermodulation products when there are multiple carriers present at the receiver's input, as previously discussed in Section 3.4.

To achieve AGC, recall that for all three detection techniques previously discussed, there is a d.c. component present in the baseband detector's output which is proportional to the input carrier level. By filtering the detector's output to remove the baseband signal f(t), the d.c. voltage thus obtained can then be used to control the gain of a selected number of RF, mixer and IF amplifiers by varying biasing conditions, e.g., by reducing the forward bias and hence the beta of a transistor to achieve a reduction in gain.

For the diode detector (Figure 3. 20), the audio baseband signal is removed from the AGC voltage line by means of the lowpass filter comprised of R_4 and C_4. The RC time constant of this lowpass filter must be longer than the period of the lowest audio frequency otherwise the AGC action will tend to suppress the audio signal itself.

The effect of automatic gain control is illustrated in

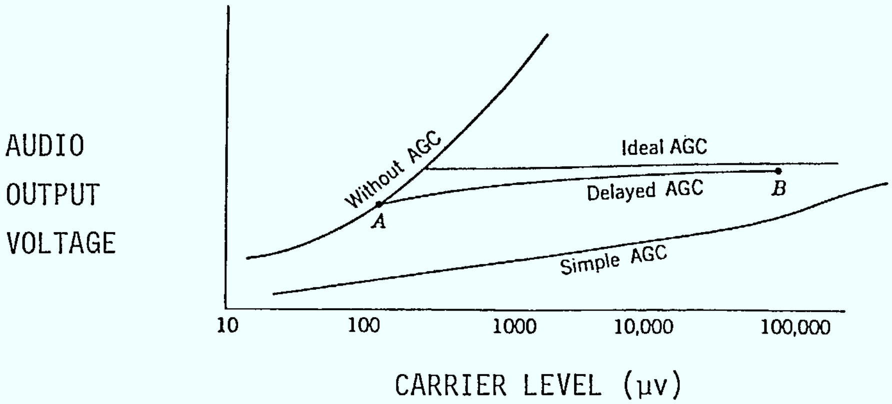


FIGURE 3.25 THE EFFECTS OF AGC ACTION

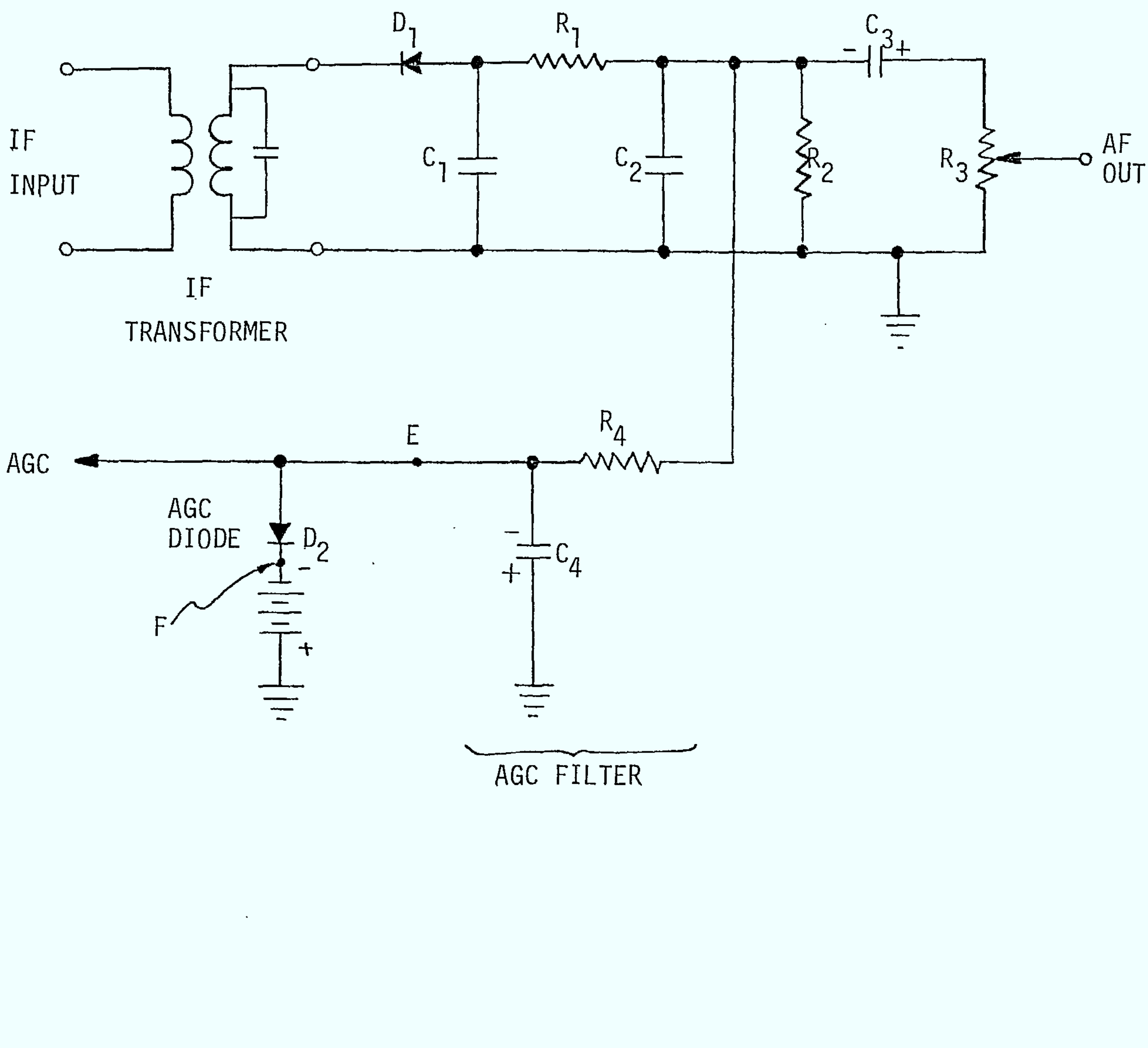


FIGURE 3.26 DELAYED AGC CIRCUIT

Figure 3.25 for different carrier levels. Without automatic gain control, the audio signal increases rapidly to a point of distortion. Ideally, the output should rise to a predetermined level and remain fixed for all higher values of carrier levels. In the simple AGC circuit, all carrier levels produce an AGC voltage and a consequent reduction in gain. Unfortunately, this simple approach decreases gain when it is most needed, i.e., at low carrier levels.

If a diode is used for automatic gain control and a separate diode for detection, the AGC diode can be biased in the manner of a gate or clamp such that it functions only when a predetermined signal level is reached (see Figure 3.26).

Alternatively, a d.c. amplifier with differential input may instead be inserted in the AGC line such that its inputs are connected to points E and F of Figure 3.26. The circuit is arranged so that amplification occurs only when the voltage at E exceeds that at F. The input to the amplifier is effectively the difference between points A and B of Figure 3.25. Amplification of the AGC voltage yields an overall receiver response that closely approaches the ideal response.

3.7.5 Choice Of Detection Technique

Envelope detection using the diode detector is employed almost universally in AM broadcast receivers for reasons of design simplicity, relatively high efficiency and low cost. There seems little reason to alter the *status quo*.

It can be shown that, for the same modulated input signal, the output of the diode detector is π times as large as that of the rectifier detector. For this reason, the diode detector is obviously preferable to the rectifier detector with which it shares a superficial resemblance.

At high input signal-to-noise (SNR) ratios, the output SNR of the diode detector is equivalent to that of the synchronous detector. However, the synchronous detector exhibits its superior noise rejection characteristics by being able to operate at low input SNR's. In contrast, the diode detector exhibits a severe degradation in output SNR, otherwise known as a *threshold effect*, due to the nonlinear action of the envelope detection process.

This capability of the synchronous detector to operate at low input SNR's suggests a definite application in coping with the problem of multipath fading in which the sky wave and ground wave originating from the same broadcast station combine destructively at distances ranging between 185 and 250 miles from the broadcast transmitter. Unfortunately, the complexity and cost involved in providing a coherent carrier reference for synchronous demodulation reflect very negatively on the use of the synchronous detector.

3.8 Design issues Relating to Spectrum Utilization Efficiency

This section draws on the discussions presented in the preceding sub-sections to address specific design issues which might possibly have an impact on AM spectrum utilization efficiency.

3.8.1 Image Rejection

As indicated in Section 3.4.3, image frequency and direct IF rejection must be provided by the RF circuitry. Since the image frequency band and the AM broadcast band partially overlap (refer to Figures 2.3 and 2.4), unrestricted channel assignment within a given broadcast area could result in image channel interference unless AM receivers are provided with sufficient image channel rejection capability.

The minimum image channel and direct IF rejection recommended by the CCIR is 30 dB (see Table 3.1). For AM receivers which utilize a ferrite-rod and no RF amplifier, this RF selectivity must be provided solely by the antenna tuned circuit.

For a receiver with a typical IF frequency of 455 kHz and a loaded antenna Q of 100, we can compute the image rejection(theoretical) provided by the single-tuned antenna circuit using equation 3.6. When the receiver is tuned to 540 kHz (the lowermost channel), the image rejection is 47.3 dB. On the other hand, the image rejection when the receiver is tuned to the uppermost channel (1600 kHz) is somewhat less, being only 39.4 dB. For the case where the band is ultimately extended to 1700 kHz, the attenuation will be 38.9 dB. Allowing a few dB's for non-ideal response, the image rejection should still exceed the minimum recommended CCIR value of 30 dB in all three cases.

It does not appear, therefore, that the image rejection provided by receivers without an RF amplifier stage presently places any fundamental restriction on the freedom with which image channels may be assigned and hence on the efficiency with which available AM channels are utilized within a given broadcast area. This being the case, the need for and the additional cost associated with an RF amplifier is probably not warranted.

All things being equal, however, a receiver with an RF amplifier stage is undoubtedly superior in performance to a receiver without one. The benefits which accrue from the use of an RF amplifier include:

1. Improved sensitivity;

2. Improved image frequency rejection;

3. Improved signal-to-noise ratio at the receiver's input;

4. Improved rejection of adjacent unwanted frequencies, i.e., better selectivity;

5. Prevention of spurious frequencies entering the mixer and heterodyning there to produce an interfering IF frequency;

6. Prevention of reradiation of the local oscillator through the receiver's antenna;

7. Better coupling of the antenna to the receiver.

For receivers located in a fairly high signal strength area, such as the central urban area of any large city, the inclusion of an RF amplifier would provide only marginal improvements in performance.

3.8.2 Selectivity

The adjacent channel selectivity of an AM receiver is provided primarily by the IF stage although it is assisted to some degree by the selectivity of the antenna and RF amplifier (if one is provided) tuned circuits (recall Figure 3.17). In the majority of receivers, the IF stage is typically comprised of two amplifier stages and three single-tuned interstage IF transformers (see Figure 3.16) which provide an overall receiver adjacent channel attenuation (ACA) of about 25 dB and a second channel attenuation (SCA)* of about 52 dB at ± 10 kHz and ± 20 kHz respectively away from the channel to which the receiver is tuned.

The adjacent channel rejection can be improved substantially by simply replacing only the second IF transformer in Figure 3.16 with a ceramic filter, such as the Murata Corporation Type CFU-455 (see Appendix A). With this circuit modification, we can reasonably expect to obtain an adjacent channel attenuation of better than 45 dB and a second adjacent channel attenuation of better than 65 dB. These values would, of course, have to be verified by a standard selectivity measurement technique.

The immediate implication of such an improvement is that adjacent channels could be more readily assigned within a given area with less concern about the possibility of adjacent channel interference and would clearly lead to more efficient usage of the available spectrum. This may alternatively be expressed in terms of a higher tolerable interference level or a reduction of the required interference protection ratio, i.e., the ratio

*For a 9 kHz channel spacing, the ACA and SCA would be slightly reduced to approximately 23 dB(±9kHz) and 49 dB (±18 kHz) respectively.

of the desired to the undesired signal strength, at the 0.5 mV/m desired signal contour.

The first IF transformer is retained to provide feed-through isolation of the RF and local oscillator into the IF stage but could also be replaced by a ceramic filter. The third IF transformer which couples the IF stage to the envelope detector is retained because of the superior diode-driving characteristic of the transformer.

The differential cost increase precipitated by the replacement of one IF transformer with a ceramic filter is approximately 75 cents for a moderately large quantity of 500 pieces. On a mass-production basis, this cost differential might be reduced to only several cents. Since the non-recurring engineering costs for accomodating a ceramic filter into an existing design is expected to be minimal and given the slight decrease in labour spent in aligning two instead of three IF transformers, the overall price impact on AM receivers is expected to be quite small. It is, of course, quite impossible for anyone but the receiver manufacturer to predict just how small,"small" really would be.

3.8.3 Sensitivity

A receiver's sensitivity is a measure of its ability to amplify weak signals which hence determines the receiver's reception range relative to any given broadcast transmitter. The CCIR recommends that AM receivers be capable of producing a standard 50 mW output power for a field strength of 5 mV/m. For a ferrite-rod antenna, this translates to an input voltage to the receiver circuitry of approximately 1 mV. For an automotive receiver which utilizes a whip antenna, the corresponding input voltage would be about 500 μV.

For a good domestic or automotive receiver which both utilize an RF amplifier stage, the sensitivity is typically in the region of 15 μV with an output signal plus noise-to-noise ratio (S+N/N) of about 20 - 25 dB. For portable and other small receivers with no RF amplifier stage, the sensitivity achieved is typically 150 μV to 300 μV with an output (S+N)/N of about 25 dB.

Since the distribution of receivers within any given broadcast coverage area is usually such that the majority of receivers lie well within the 5 mV/m signal level contour of the transmitter, only a small percentage of small portable receivers located in the fringe areas of reception will tend to be affected by limited sensitivity.

As indicated earlier, receiver sensitivity and signal-to-noise performance can be improved with the addition of an RF amplifier stage but improving the sensitivity to a level below 100 μV must be weighed against the following:

1. The cost of adding an RF stage

2. The number and type of receivers affected

3. The fact that the level of environmental and/or man-made noise at 1 MHz is in region of 20 to 200 μV/m (i.e., approximately 5 to 50 mV input voltage level for receivers with ferrite-rod antennas)

4. The increased susceptibility to co-channel interference.

On the whole, it is felt that improving the sensitivity of AM receivers, currently without an RF stage, to below 100 μV is not warranted.

3.8.4 Overload Levels

The overload level for AM receivers is defined in terms of the RF input level required to increase the audio distortion to 10% at a 30% modulation depth. It is not known what the present overload performances of AM receivers are but for AM radio integrated circuit (IC) receivers, RF input levels of greater than 100 mV or, in some cases, greater than 1V can be accomodated. For ferrite-rod antennas, these input voltage levels would correspond to field strengths of about 400 mV/m and 4 V/m respectively; for whip antennas, the corresponding field strenghts would be 1 V/m and 10 V/m respectively. Thus, for AM IC receivers at least, this level of performance is sufficient to accomodate field strengths which are only encountered in close proximity to the broadcast transmitting antenna.

To avoid blanketing radio receivers, Broadcast Procedure 1, Rule 2 establishes a maximum field strength contour of 1 V/m within which only a minimum of buildings and population is allowed to be enclosed. In addition, Rule 2 also stipulates the maximum population that may be enclosed by the 250 mV/m contour as a function of broadcast station power; furthermore, this enclosed population must not be greater than one-third of the total population within the principal center to be served. It would seem, therefore, that gain compression and the resulting audio distortion due to RF signal overload should not be a problem for the majority of receivers which have no RF stage. For those fitted with an RF amplifier, AGC is usually applied to both the RF stage as well as the first IF amplifier stage to cope with abnormal signal level conditions.

3.8.5 Intermodulation

Intermodulation (IM) distortion products are generated when two or more signals are amplified or mixed in a nonlinear device. For a large number of AM receivers which utilize a ferrite-rod antenna and no RF stage, poor RF selectivity by the antenna tuned-circuit can result in unwanted adjacent signals combining with a desired signal in the mixer stage of the receiver to produce IM products which may fall within the passband of the IF stage. The IM products which are usually of greatest concern are the third-order IM products which tend to fall within the passband of even moderate-bandwidth amplifiers.

Interference due to IM products may be minimized by:

1. Improving RF selectivity by providing an RF amplifier stage (with AGC capability)

2. Improving the 1 dB conversion compression point and third-order intercept point of the mixer stage through improved semiconductor devices such as MOSFET transistors which typically exhibit high dynamic range, excellent isolation between the LO and RF signal and high third-order intercept points. (FET's, in general, exhibit a square-law transconductance response which lacks higher-order power terms).

3. Utilizing balanced mixer structures such as differential transistor pairs.

4. Employing circuit techniques such as negative feedback.

Some presently available AM radio IC's (e.g., Signetics TCA440) employ balanced mixers that result in the generation of fewer harmonic mixing products and whistling points.

In the particular case of the whistle which occurs at 910 kHz due to the second harmonic of the signal mixing with the LO frequency of 1,365 kHz, partial relief may be obtained by slightly detuning the IF but the whistle will still be present. A more satisfactory solution would be to employ a balanced mixer structure, as mentioned above, which would tend to suppress even harmonic products at the mixer's output.

3.8.6 Standardization of The LO Frequency Range

With the present tuning range for the AM band at 540 to 1600 kHz and an IF frequency of 455 kHz (typically), the local oscillator tuning range becomes 995 to 2055 kHz and the image frequency band, 1450 to 2510 kHz. Since the separation between a desired signal frequency and its image is equal to twice the IF frequency or 910 kHz in this case, a single-tuned circuit will provide sufficient image rejection, as shown earlier in Section 3.8.1.

An IF frequency which is chosen somewhere in the range 440 to 470 kHz would thus result in a corresponding image frequency separation of between 880 and 940 kHz. By applying equation 3.6, the image frequency rejection values obtained for a receiver without an RF stage and with a loaded antenna Q of 100 are as follows:

(a) IF = 440 kHz:

Receiver tuned to (kHz)	Image Frequency (kHz)	Image Rejection (dB) (Q_L = 100)
540	1420	47.0
1600	2440	38.7
1700	2540	38.3

(b) IF = 470 kHz

Receiver tuned to (kHz)	Image Frequency (kHz)	Image Rejection (dB) (Q_L = 100)
540	1480	47.5
1600	2540	39.6
1700	2640	39.2

It is clear from the above results that, even allowing a couple of dB's for non-ideal implementation, the image rejection obtained for both the present and future extended AM band still exceeds the minimum recommended CCIR value of 30 dB for either extreme of IF frequency. We therefore conclude that the actual choice of IF frequency, within the range 440 to 470 kHz, does not present any limitations on the freedom with which the image channels may be assigned.

3.8.7 Increased Tuning Range

For an IF of 455 kHz, the LO tuning range is 995 to 2055 kHz which gives a maximum-to-minimum frequency ratio of

2.06:1 and a corresponding minimum-to-maximum capacitance ratio of 1:4.26. Since the capacitance ratio typically obtainable is 10:1 for a tuning capacitor, a value of 4.26:1 can thus be easily accomodated. For the case where the AM band is extended from 1600 to 1700 kHz, the new frequency ratio would be 2.16:1 and the new capacitance ratio would be 4.69:1, which is still well within the tuning capacitor's capability.

For the RF tuned circuit, the present maximum-to-minimum frequency ratio is 1600: 540, i.e., 2.96:1 and therefore the required tuning capacitance ratio is 8.8:1. For the extended frequency band, the new frequency ratio becomes 3.15:1 and the required tuning capacitance ratio, 9.9:1. This capacitance ratio is very close to the present limit of 10:1 but should not present a problem.

As an alternative, one might perhaps consider the use of varactor tuning diodes* which can provide capacitance ratios in excess of 14:1 and, in some cases, even as high as 28:1.

3.8.8 Digital Tuning

In the past, digital tuning has mainly been associated with complex avionic systems. Many recent developments have stimulated its application in the consumer market including:

1. FCC legislation [3] on uniform methods of selecting and displaying VHF and UHF TV channels;

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*To facilitate the use of digital techniques in radio tuning, Philips Electronics has developed a dual variable-capacitance diode (BB212) specifically for AM band tuning.

[3]FCC Part 15 Regulations, Section 15.68, Paragraph b3.

2. The manufacturers' desires for minimizing internal adjustments and alignments;

3. The considerable and ever-growing popularity of Citizen Band (CB) radio with its many available and impending new channels which make the digital frequency synthesizer approach the only economical one;

4. The consumer's desire for simpler and more effective controls.

The availability of MOS LSI frequency synthesizers, high-speed ECL counters and varactor tuners now make the digital tuning solution possible.

Operational features of a digitally tuned receiver include:

- Precise tuning of station frequency
- Display of exact receiver frequency
- Keyboard entry of station frequency
- Storage of multiple stations in memory
- Pushbutton up/down scan through the band
- Pushbutton search (stop on next station)
- Power-on to the last station selected
- Provision for time-of-day clock
- Ability to address the synthesizer

There are two basic approaches to the implementation of digital tuning for AM radios:

Method A: Fully programmable (or random access)frequency selection. This is usually provided by using a programmable divider and reference oscillator phase-locked loop (PLL) technique (see, for example, Figure 3.27);

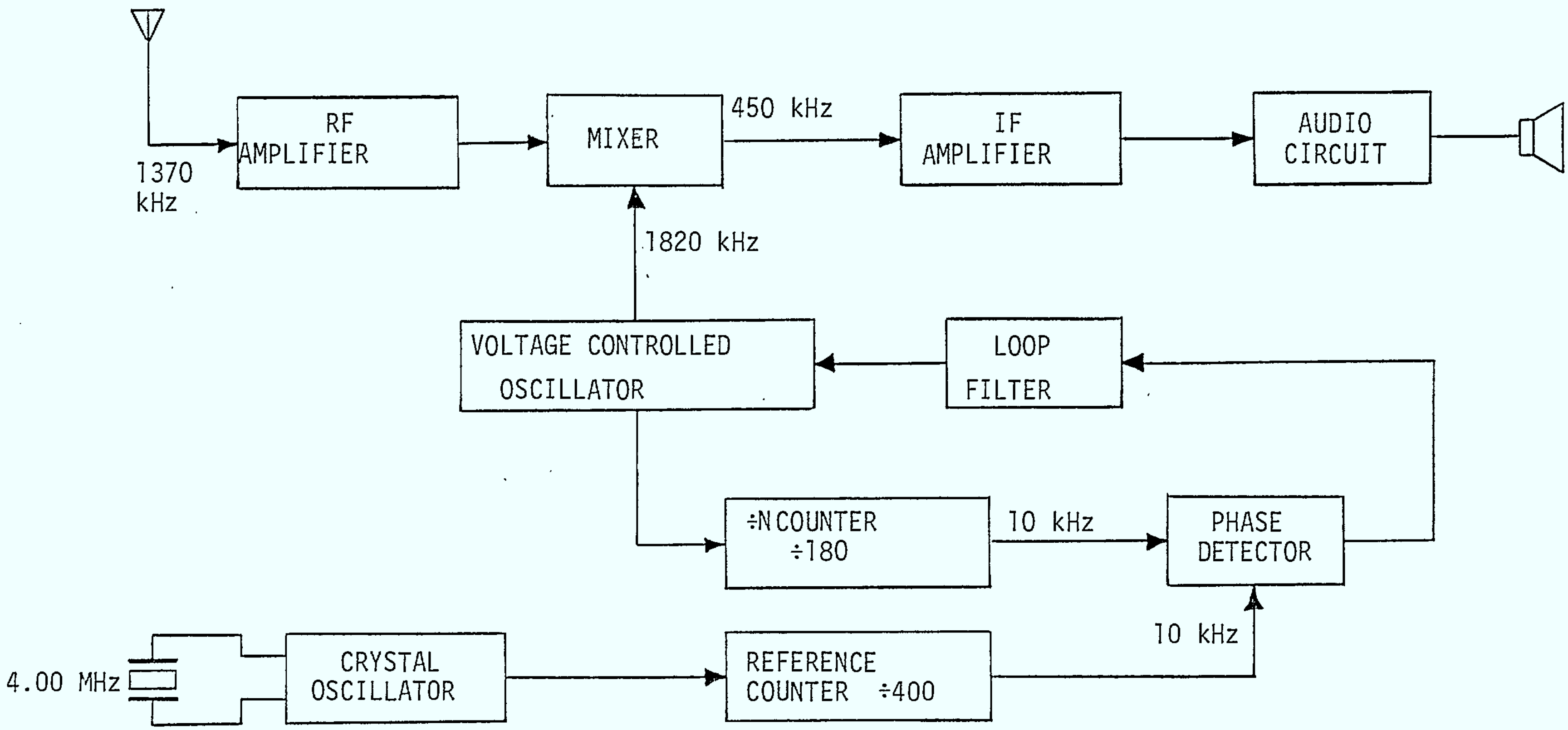


FIGURE 3.27 AN EXAMPLE OF AN AM PHASE-LOCKED LOOP (PLL) SYNTHESIZER

Method B: Coarse mechanical tuning with a "fine tune" to lock the receiver LO to the nearest multiple of the reference frequency.

Method A requires a wide-range voltage-controlled oscillator (VCO) but can otherwise be fully digital in operation with keyboard entry of the desired station frequency. Method B requires a manually variable "coarse-tune" control (e.g., a potentiometer/varactor tuning diode combination or a conventional mechanically-tunable capacitor). The "fine-tune" control could be exercised by simply modifying an existing tuning system.

It has been suggested [4] that the continuous-rotation tuning control is probably the most acceptable from the user's viewpoint and, hence, Method B would seem the better choice especially for low-cost applications.

In most PLL synthesizer implementation for AM applications, the IF frequency chosen for the system is 450 kHz. This is necessitated by the fact that the LO must lock to multiples of the channel spacing being used, i.e., 10 kHz in the case of AM. As illustrated in Figure 3.27, the synthesizer's reference frequency (10 kHz) is derived by dividing the output frequency of a crystal oscillator. It is rather fortuitous that the choice of a 450 kHz IF also allows the use of a 9 kHz channel spacing by a simple change of the crystal.

While no costs are available for the implementation of digital tuning in AM receivers, we can, nevertheless, surmise that such a feature would probably be incorporated

[4]Andrew C. Tickle, "Channel Selection Methods for Digital Tuning", IEEE Trans. on Consumer Electronic, Vol CE-23, No. 3, August 1977, pp. 195 - 199.

only in the premium-cost receivers, i.e., the console-type AM/FM receivers and some AM/FM automotive receivers. Digital tuning may eventually be incorporated into the medium-cost hand-carried AM receivers as the cost of such systems continue to decrease but is unlikely to be applied to the low-cost portable pocket radios.

In summary, Table 3.2 provides a comparison of mechanical tuning versus synthesized tuning for various design features that should be considered for AM receiver tuning application.

TABLE 3.2

SUMMARY OF TUNING METHOD CONSIDERATIONS FOR AM BROADCAST RECEIVERS

FEATURE	MECHANICAL TUNING	SYNTHESIZED TUNING
•Continuous rotation	Inherent	Disc with slot/holes interrupting a light beam
•Keyboard frequency entry	Not practical	Inherent
•Automatic Scan	Difficult - need motor drive	Yes
•Memory of last station tuned	Inherent	Continuous application of power to parts of the circuits. Special ICs with non-volatile storage
•Display of tuned frequency	- Mechanical Dial - Counter with IF offset.	- Thumbwheels - Register/counter with IF offset
•Accuracy (Drift)	Can be improved with varicap fine-tune and locking to reference harmonic.	Inherent
•Spurious responses or RFI generation	Little or not trouble, except for multi-plexed LED display	Need careful design Multiplexed LED-RFI
•Integration	Except for mechanically variable element.	Except frequency data input device.
•Cost	- Low - Intermediate with Lock and Display	- Intermediate, probably more than mechanical with Lock and Display

4.0 CONCLUSIONS

With the success of Canada's submission to the recent World Administrative Radio Conference (WARC) held in Geneva, the extension of the upper frequency limit of the AM broadcast band from 1600 kHz to 1700 kHz is now imminent. To examine the impact this and other factors might have on the AM broadcast receiver and the manner in which the AM spectrum might be used more efficiently, a "broad-brush" treatment of the technical aspects surrounding AM broadcasting has been presented, followed by a fairly detailed examination of the receiver itself as an essential step in understanding what improvements might or might not be achieved in making certain design modifications.

Unfortunately, the attempts to evaluate the effect, both technically and cost-wise, of certain design modifications to the broadcast receiver has generally been hampered by the general lack of any concerted body of information on the actual performance characteristics of modern transistorized AM receivers. This is partly due to the fact that there are no rigid technical specifications that AM receivers are required to meet. The only specifications relating to the performance of AM receivers are contained in CCIR's Recommendation 415 which, in fact, only recommends the minimum performance specifications that receivers should meet. This is perfectly understandable, however, since the main objective of Recommendation 415 is to encourage the large-scale production of low-cost sound-broadcasting receivers for worldwide distribution.

With respect to the actual design modifications, there is unquestionably a lot that can be done to improve the performance of AM receivers but the important question is: "At what cost?". The first desirable, but not necessarily essential, modification would be to provide an RF amplifier stage for all receivers, including the portable pocket receivers which are currently sold at prices ranging from $5 to $10. The immediate benefits of adding an RF stage, as discussed in Section 3.8, would include improved sensitivity, selectivity, image rejection and signal-to-noise performance.

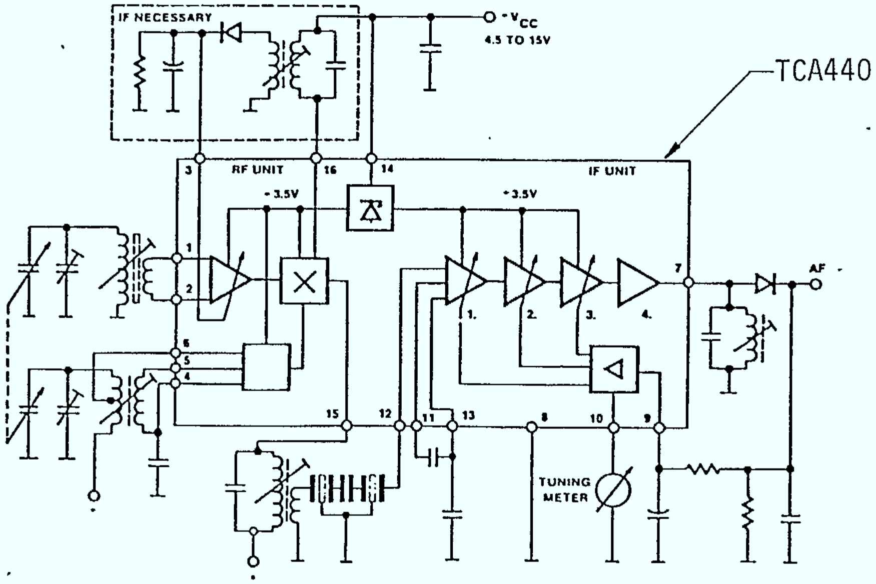
A second modification which would be very beneficial but yet is expected to have a minimal impact on receiver cost, is that of replacing one of the single-tuned interstage transformers in the IF stage of the receiver with a modern ceramic filter. The adjacent channel rejection of the receiver would be significantly improved by this modification with consequent effect on the present allowable interference levels and required spacings between broadcast transmitters. The net effect would be to allow more efficient utilization of AM channels, which was one of the prime objectives of this study. With the possibility of channel spacings being decreased from 10 kHz to 9 kHz, improving the adjacent channel rejection and overall selectivity of AM receivers will be essential.

Other possible improvements which could be implemented include digital synthesis and tuning for the local oscillator; the use of balanced mixers or improved mixing devices such as MOSFET transistors; the use of synchronous detection instead of envelope detection for improved audio signal-to-noise performance particularly at low input signal-to-noise ratios, as in multipath fade conditions.

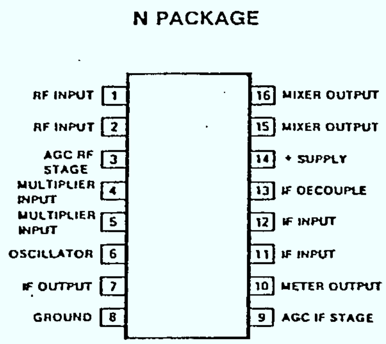
A more satisfactory approach to the whole question of receiver modifications would, in fact, be to replace nearly all of the active discrete components with a single AM receiver integrated circuit (IC) which, judging from the manufacturers' performance data, generally provide far superior performance to the discrete component approach. One prime example is the Signetics TCA440 AM receiver IC (see Figure 4.1 for a block diagram and pin configuration for the IC and Figure 4.2 for an example of a radio receiver based on the IC) which is quite inexpensive at about $2.32 each in a quantity of only 500. Less "exotic" AM receiver IC's, such as National Semiconductor's LM1820 and RCA's CA3088E, are also available at OEM* prices ranging from $2 to $3 each in quantity 500. In mass-production quantities, AM receivers IC's should be very competitive with the discrete approach. It appears, however, that the general tendency in radio receivers today is to combine the AM function with FM, using widely available AM/FM receiver IC's, while the discrete approach is still applied to the strictly AM receivers.

No attempts have been made to "guesstimate" what cost impact various design modifications might have on the various classes of AM receivers (low cost, medium cost and premium) because such estimates would obviously be fraught with inaccuracy; rather, component costs for a moderately large quantity of 500 have been presented when available. Generally speaking, however, it is believed that any changes that will improve the performance of AM receivers will find acceptance among consumers, providing, of course, cost increases are not disproportionately large. Ultimately, this decision will be made by the consumer.

*Original Equipment Manufacturers



(a) Block Diagram



(b) Pin Configuration

FIGURE 4.1 SIGNETICS TCA440 AM RECEIVER IC
(a) BLOCK DIAGRAM
(b) PIN CONFIGURATION

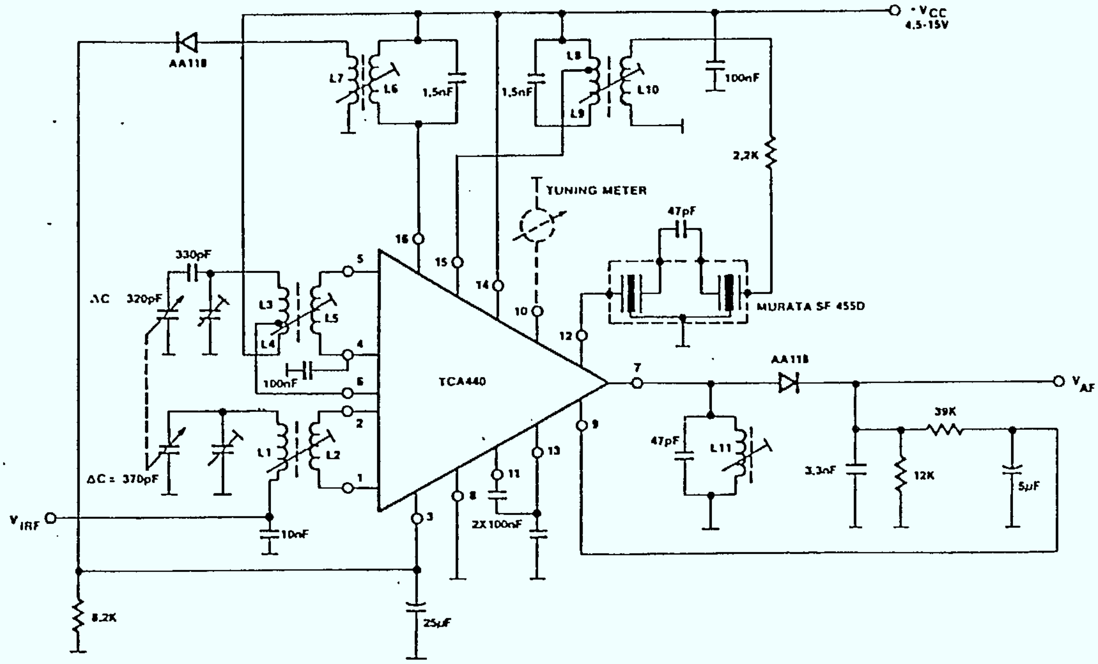


FIGURE 4.2 AM RECEIVER* USING THE TCA440 INTEGRATED CIRCUIT

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* Reproduced from Signetics Analog Data Manual (1979), p. 268.

REFERENCES

[1] CCIR XIIIth Plenary Assembly, Geneva, 1974, Greenbook Vol. X, pp. 119 - 123.

[2] H. J. Laurent and C. A. B. Carvalho, "Ferrite Antennas for AM Broadcast Receivers", Application Note, Bendix Radio Division of the Bendix Corporation.

[3] FCC Part 15 Regulations, Section 15.68, Paragraph b3

[4] Andrew C. Tickle, "Channel Selection Methods for Digital Tuning", IEEE Trans. on Consumer Electronics, Vol. CE-23, No.3, August 1977, pp. 195 - 199.

APPENDIX A

IF FILTERS FOR AM RECEIVER APPLICATIONS

This appendix examines some of the common passive IF filters available today which might or might not be suitable for application in AM broadcast receivers. The filter types to be considered here are:

1. The LC single-tuned or double-tuned IF transformer

2. The crystal filter

3. The mechanical filter

4. The Surface Acoustic Wave (SAW) filter

5. The ceramic filter

A.1 Definition of Filter Response

Critical to the understanding of filter behaviour is a definition of the vocabulary of the most frequently used terms and familiarity with the typical filter amplitude frequency response curve (Figure A.1):

(a) Center Frequency (fo) - The arithmetic mean between the high and low cut-off frequencies of a filter.

(b) Bandwidth (BW) - The difference between two limiting frequencies at a specified attenuation level.

(c) Attenuation - Reduction of signal level in transmission through a filter. (Attenuation is usually expressed in decibels (dB).

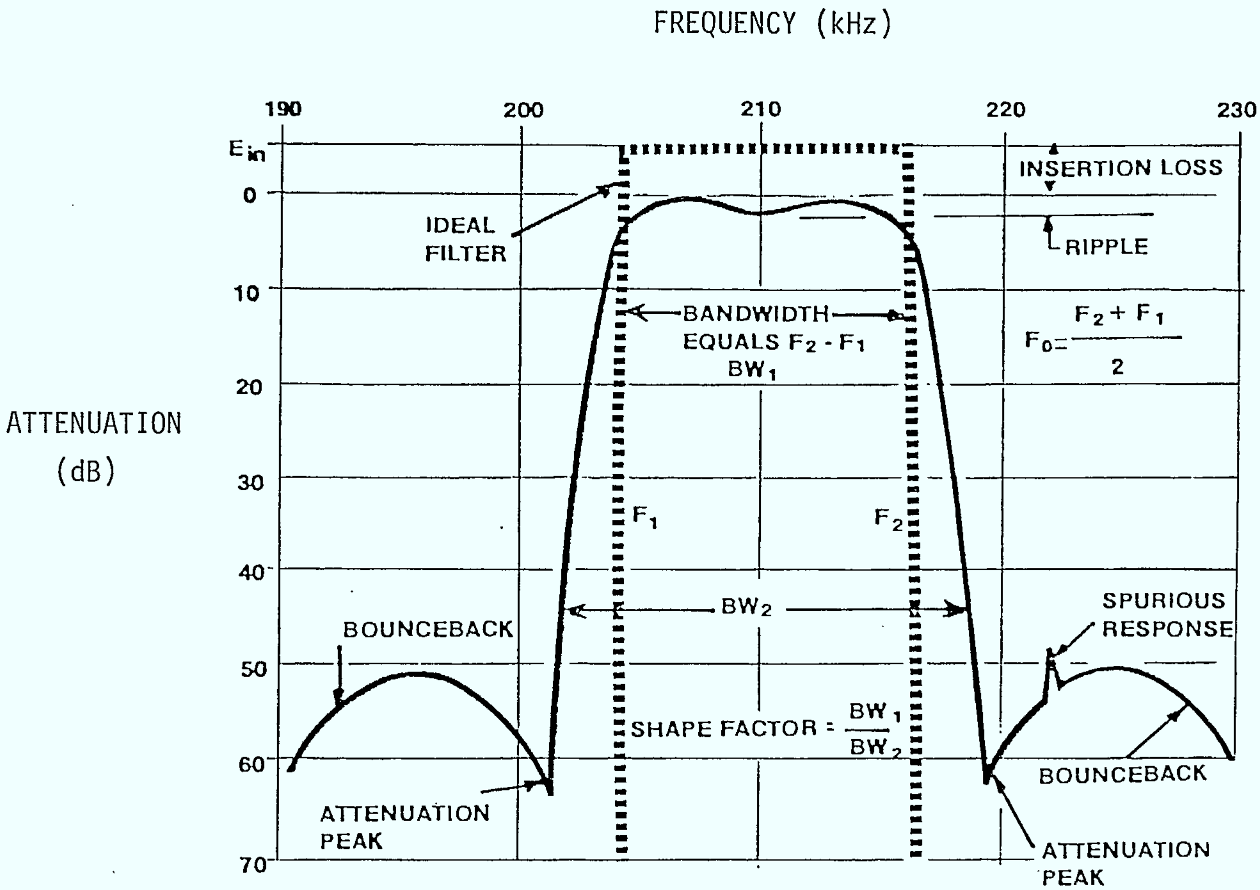


FIGURE A.1 FILTER RESPONSE PARAMETERS

(d) Decibel - Unit that expresses the ratio between two powers, two voltages or two currents

$$\left(10 \text{ Log}\frac{P_1}{P_2}, 20 \text{ Log}\frac{V_1}{V_2} \text{ or } 20 \text{ Log}\frac{I_1}{I_2}\right)$$

(e) Shape Factor - Ratio of bandwidths at two different levels of attenuation.

(f) Ripple - The wavelike response in the passband of a filter (expressed in dB). The maximum ripple is the excursion from the highest peak to the lowest valley.

(g) Insert Loss - Power loss of the filter in the passband (expressed in dB). Zero dB reference is the point of maximum output of the filter unless it is specified otherwise.

$$\text{Insertion Loss} = 10 \text{ Log} \frac{P \text{ in}}{P \text{ out}}$$

(h) Source impedance - (input termination) - The output impedance of the circuit that drives the filter.

(i) Load Impedance - (output termination) - The impedance that must be connected to the output terminals of the filter in order to achieve the proper response.

A.2 Single-Tuned/Double-Tuned IF Transformer

In the majority of AM broadcast receivers, the intermediate frequency (IF) amplifier is comprised of a cascade of a number of stages whose frequency response is determined by

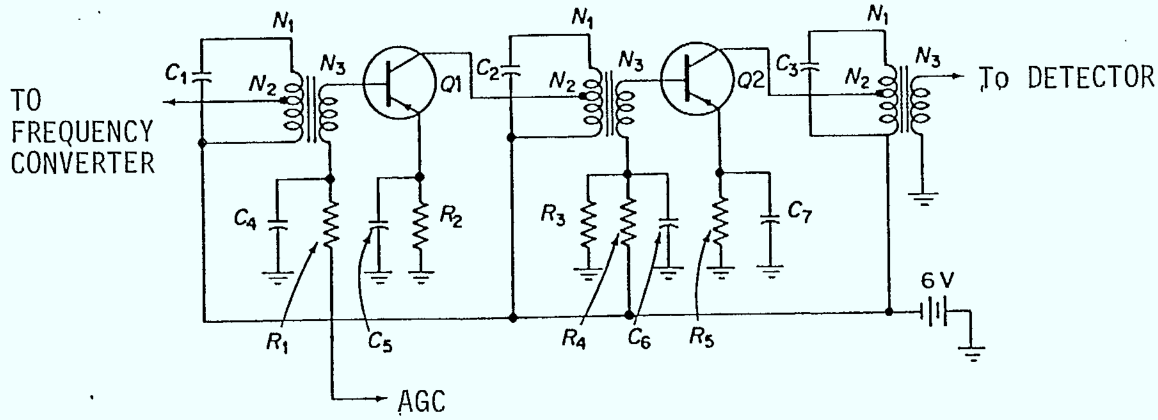


FIGURE A.2 SINGLE-TUNED IF AMPLIFIER STAGES

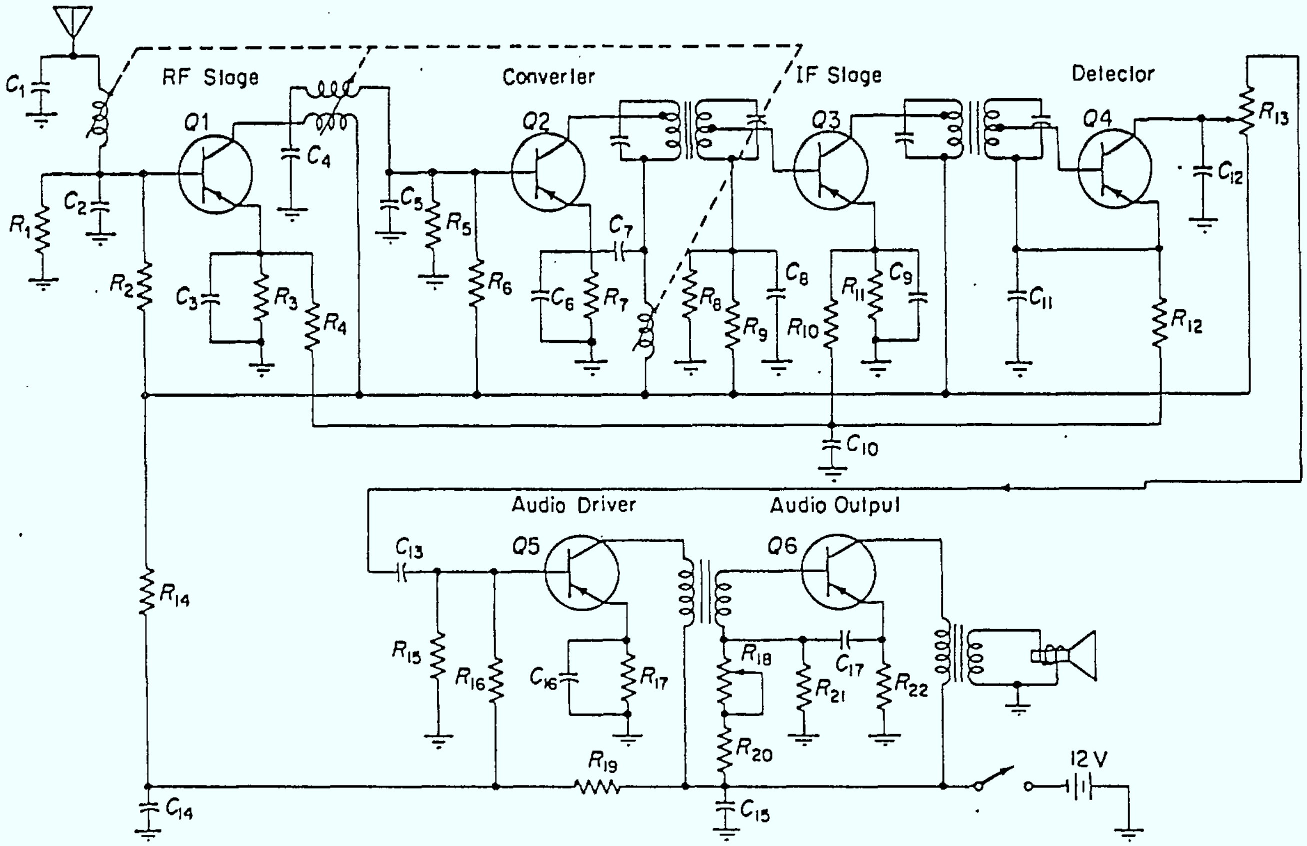


FIGURE A.3 AM RECEIVER WITH A SINGLE- DOUBLE-TUNED IF STAGE tuned interstage IF transformers. The interstage IF transformer may either be single-tuned as, for example, in Figure A.2 or double-tuned (Figure A.3).

A.2.1 Description

The IF transformers commonly consist of one or more resonant circuits which are usually inductively coupled together and mounted in a metal shield.

The transformers are tuned by means of small, parallel-connected capacitors. The capacitors are variable in some cases and in others, the capacitors are fixed and the winding is tuned by varying the position of the slug core.

The windings of the transformer consist of small, flat, universal-wound pies mounted on either an insulated core or on a powdered-iron core. The iron-core transformers generally have somewhat more gain and better selectivity than equivalent air-core units.

A.2.2 Single-Tuned Transformer

Single-tuned circuits are commonly used in the IF stage of many AM receivers because of the higher voltage gain which may be achieved (as compared to double-tuned transformers). The bandpass response of the IF stage is achieved by cascading a number of synchronously tuned (i.e., tuned to the same frequency) amplifier stages. The overall bandwidth then achieved is always smaller than that of a single stage but the skirt selectivity is greatly improved.

Bandwidth shrinkage may be avoided, however, by ensuring that each stage is tuned to a different frequency. For a flat passband response, the individual stages are geometrically balanced about the center frequency. Stagger-tuning is rarely, if ever, used in AM broadcast receivers.

For the single-tuned stage (Figure A.4), the amplitude/frequency response is given by:

$$A(j\omega) = -A_r \frac{1}{1 + j\,Q_L\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)} \tag{A.1}$$

where

Q_L = loaded Q of the tuned circuit (>10)

ω_o = resonant frequency of the tuned circuit = $\frac{1}{\sqrt{LC}}$

ω = frequency variable

A_r = midband gain equal to g_m times the midband impedance level.

For an n-stage amplifier with n interstages:

$$A_T = A^n(j\omega) = A_r^n \left[1 + \left(\frac{\omega^2 - \omega_o^2}{B\omega}\right)^2\right]^{-\frac{n}{2}} \tag{A.2}$$

where

$B = \frac{\omega_o}{Q_L}$ = single stage bandwidth

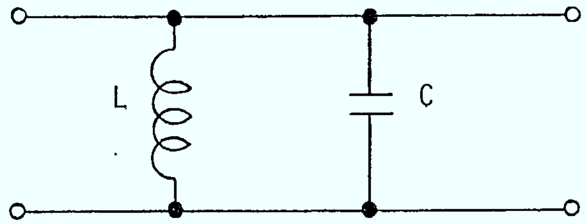


FIGURE A.4 SINGLE-TUNED TRANSFORMER

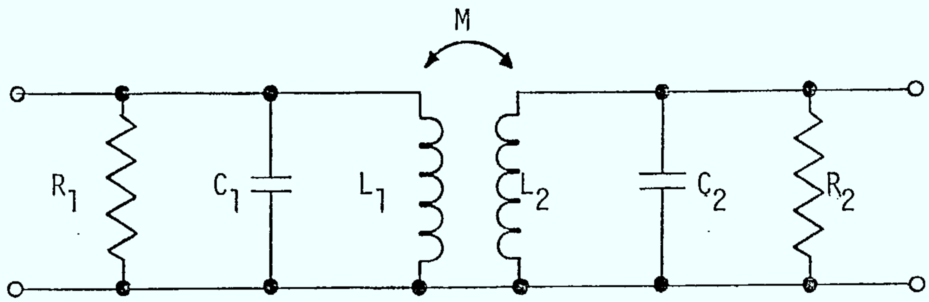


FIGURE A.5 DOUBLE-TUNED TRANSFORMER

n = number of stages

ω_o = center frequencies

Q_L = loaded Q

A.2.3 Double-Tuned Transformer

The double-tuned IF transformer consists of a primary and a secondary which are both tuned to the same frequency and coupled inductively to a degree dependent on the desired shape of the amplitude/frequency response curve. The amplitude/frequency response for a single double-tuned stage (see Figure A.5) is given by:

$$A(j\omega) = \frac{g_m k}{C_1 C_2 (1-k^2)\sqrt{L_1 L_2}} \frac{j\omega}{\omega^4 - ja_1\omega^3 - a_2\omega^2 + ja_3\omega + a_4} \qquad (A.3)$$

where

$$a_1 = \omega_r\left(\frac{1}{Q_1} + \frac{1}{Q_2}\right)$$

$$a_2 = \frac{\omega_r^2}{Q_1 Q_2} + \frac{1}{1 - k^2}(\omega_1^{\,2} + \omega_2^{\,2})$$

$$a_3 = \frac{\omega_r}{1 - k^2}\left(\frac{\omega_2^{\,2}}{Q_1} + \frac{\omega_1^{\,2}}{Q_2}\right)$$

$$a_4 = \frac{\omega_1^{\,2}\,\omega_2^{\,2}}{1 - k^2}$$

The circuit parameters are:

R_1 = total resistance primary side

C_1 = total capacitance primary side

L_1 = total inductance primary side

R_2 = total resistance secondary side

C_2 = total capacitance secondary side

L_2 = total inductance secondary side

M = mutual inductance = $k\sqrt{L_1 L_2}$

k = coefficient of coupling

ω_r = resonant frequency of amplifier

$\omega_1 = 1/\sqrt{L_1 C_1}$

$\omega_2 = 1/\sqrt{L_2 C_2}$

Q_1 = primary Q at $\omega_r = \omega_r C_1 R_1$

Q_2 = secondary Q at $\omega_r = \omega_r C_2 R_2$

g_m = transconductance of active device at midband frequency

If $\omega_1 = \omega_2 = \omega_0$, that is, the primary and secondary are tuned to the same frequency, then

$$\omega_r = \frac{\omega_0}{\sqrt{1 - k^2}} \tag{A.4}$$

is the resonant frequency of the amplifier and

$$A(j\omega_r) = \frac{+jkg_m\sqrt{R_1 R_2}}{\sqrt{Q_1 Q_2}(k^2 + 1/Q_1 Q_2)} \tag{A.5}$$

is the gain at this resonant frequency. For maximum gain,

$$k_c = \frac{1}{\sqrt{Q_1 Q_2}} = \text{critical coupling} \tag{A.6}$$

and for maximum flatness,

$$k_T = \sqrt{\frac{1}{2}\left(\frac{1}{Q_1{}^2} + \frac{1}{Q_2{}^2}\right)} = \text{transitional coupling} \tag{A.7}$$

if k is increased beyond k_T, a double-humped response is obtained.

Figure A.6 illustrates the selectivity responses for a two-stage, double-tuned circuit, where $k_T = k_c$, for various degrees of coupling (k/k_c).

A.2.4 Bandwidth

For an n-stage, single-tuned IF amplifier circuit, the overall bandwidth, B_n^s, of the circuit is given by:

$$B_n^s = B_s\sqrt{2^{\frac{1}{n}} - 1} \tag{A.8}$$

where B_s = single-tuned amplifier bandwidth.

For an n-stage amplifier, having equal Q circuits with double-tuned interstages of bandwidth B_d, the overall bandwidth, B_n^d, is given by:

$$B_n^d = B_d\left(2^{\frac{1}{n}} - 1\right)^{\frac{1}{4}} \tag{A.9}$$

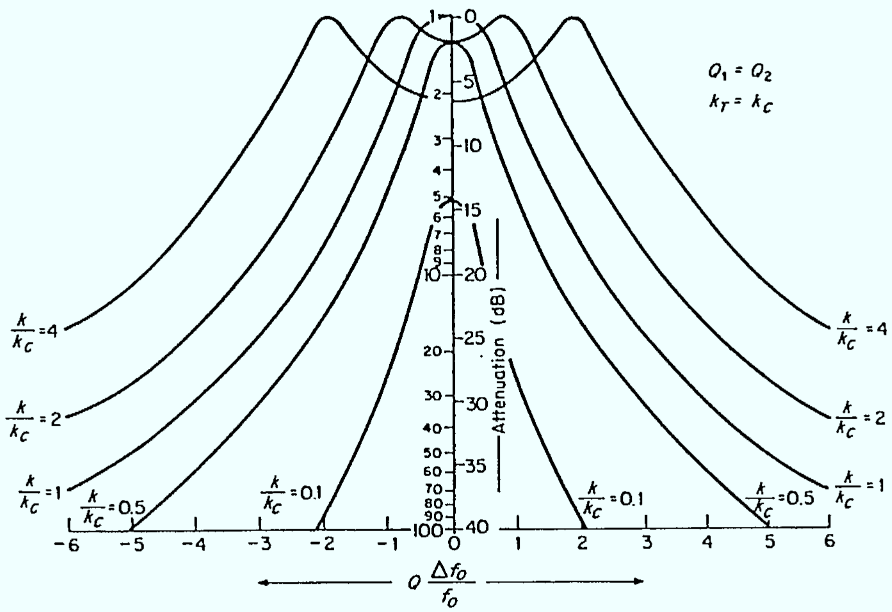
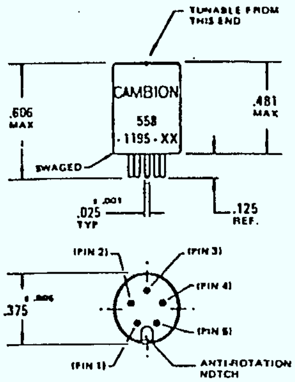


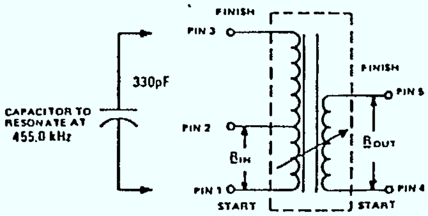
FIGURE A.6 SELECTIVITY CURVES FOR A TWO STAGE, DOUBLE-TUNED CIRCUIT

NO. OF STAGES	RELATIVE VALUES OF 3-dB BANDWIDTH		SHAPE FACTORS (BW_{60}/BW_6)	
	SINGLE-TUNED	DOUBLE-TUNED*	SINGLE-TUNED	DOUBLE-TUNED
1	1.00	1.00	577	23.9
2	0.65	0.80	33	5.65
3	0.51	0.71	13	3.59
4	0.44	0.66	8.6	2.94
6	0.35	0.59	5.9	2.43
8	0.30	0.55	5.0	
10	0.27	0.52	4.5	

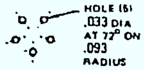
*Based upon identical primary and secondary circuits critically coupled

TABLE A.1 RELATIVE VALUES OF 3 dB BANDWIDTH AND SHAPE FACTORS FOR SINGLE- AND DOUBLE-TUNED CIRCUITS





RECOMMENDED
MOUNTING HOLES
(TOL ±.003)



OPERATING CHARACTERISTICS

BASIC PART No.		IMPEDANCE RATIO (Ω) R in	R out	FREQUENCY (kHz)	Q Min. UNLOADED	Q LOADED	INSERTION LOSS (dB)	SRF Min. (MHz)
558 - 1195	-01	25.0k	500	455.0	135	34 - 42	2.25	3.6
	-02	71.0k	1.0k	455.0	135	34 - 42	1.90	3.6
	-03	1.0k	500	455.0	135	34 - 42	2.46	3.6
	-04	10.0k	1.0k	455.0	135	34 - 42	2.25	3.6
	-05	20.0k	5.0k	455.0	135	34 - 42	2.26	3.6

558 - 1195 Transformer is identical, externally to 533 - 1181 Shielded Coil Form. Windings are varnish impregnated and ferrite components are moisture proofed.

CUP CORE MATERIAL IS: - 01 thru - 05 High Q Ferrite

FIGURE A.7 455 kHz IF TRANSFORMER CHARACTERISTICS
(CAMBRIDGE THERMIONIC CORPORATION)

Table A.1 summarizes the relative 3 dB bandwidths and shape factors (60 dB BW: 6 dB BW) for cascaded single- and double-tuned amplifier stages.

A.2.5 Suitability for AM Receiver Applications

IF transformers have enjoyed widespread usage in AM receivers mainly because of their relatively straight forward manufacturing requirements which routinely yield in-circuit performances close to predicted values.

Double-tuned IF transformers can provide flatter pass-band response and steeper skirt selectivity than the single-tuned IF transformer, as shown above, but they are more expensive and require more effort for proper alignment. Where limited selectivity is sufficient, single-tuned transformers are therefore used. The main disadvantages to the use of IF transformers for interstage coupling is the time required to properly align the IF stage response and the limited selectivity that can be provided.

Figure A.7 provides the dimensions and operating characteristics of one type of high-performance 455 kHz IF transformer manufactured by the Cambridge Thermionic Corporation for communication receivers. The typical cost of IF transformers for AM receiver applications is approximately $2.61 (quantity 500).

A.3 Crystal Filters

A.3.1 General

Crystal bandpass filters can be divided into two basic categories: the narrowband filters in which the filtering elements consist of crystals and capacitors only; and the wideband filters which utilize inductances either in series or in parallel with the crystal units. Narrowband crystal filters can be designed with bandwidths as narrow as .005% of the center frequency and up to .7% of the center frequency. The lower limit is determined primarily by the Q of the filter element while the higher limit is established by the zero-pole spacing of the particular crystal element being used.

Bandwidths for the wideband category extend from .6% of center frequency to 10% of center frequency. Again the lower limit is determined primarily by the Q of the inductances used in conjunction with the crystals. Theoretically, the upper limit depends on the zero-pole spacing of the crystal element. However, in practice it is usually determined by limitations in coil design. The wider bandwidth filters require extremely high inductances with little distributed capacity. In some instances, primarily in the frequency range where AT cut crystals are used, the upper limit is set by spurious modes in the crystals (see Figure A.8).

The shape factor of both narrowband and wideband filters depends upon the type of design and the number of sections being used in the particular filter (see Figure A.9).

Filters may be made up in two different ways. One is to cascade two or more simple sections until the desired shape factor is achieved. This method is desirable when the number of sections being cascaded is not excessive. It is easy to align, similarity of components simplifies production, and it provides maximum attenuation of spurious responses in the reject bands.

PERCENT BANDWIDTH

(3 dB)

$$\% = \frac{BW_3}{f_0} \times 100$$

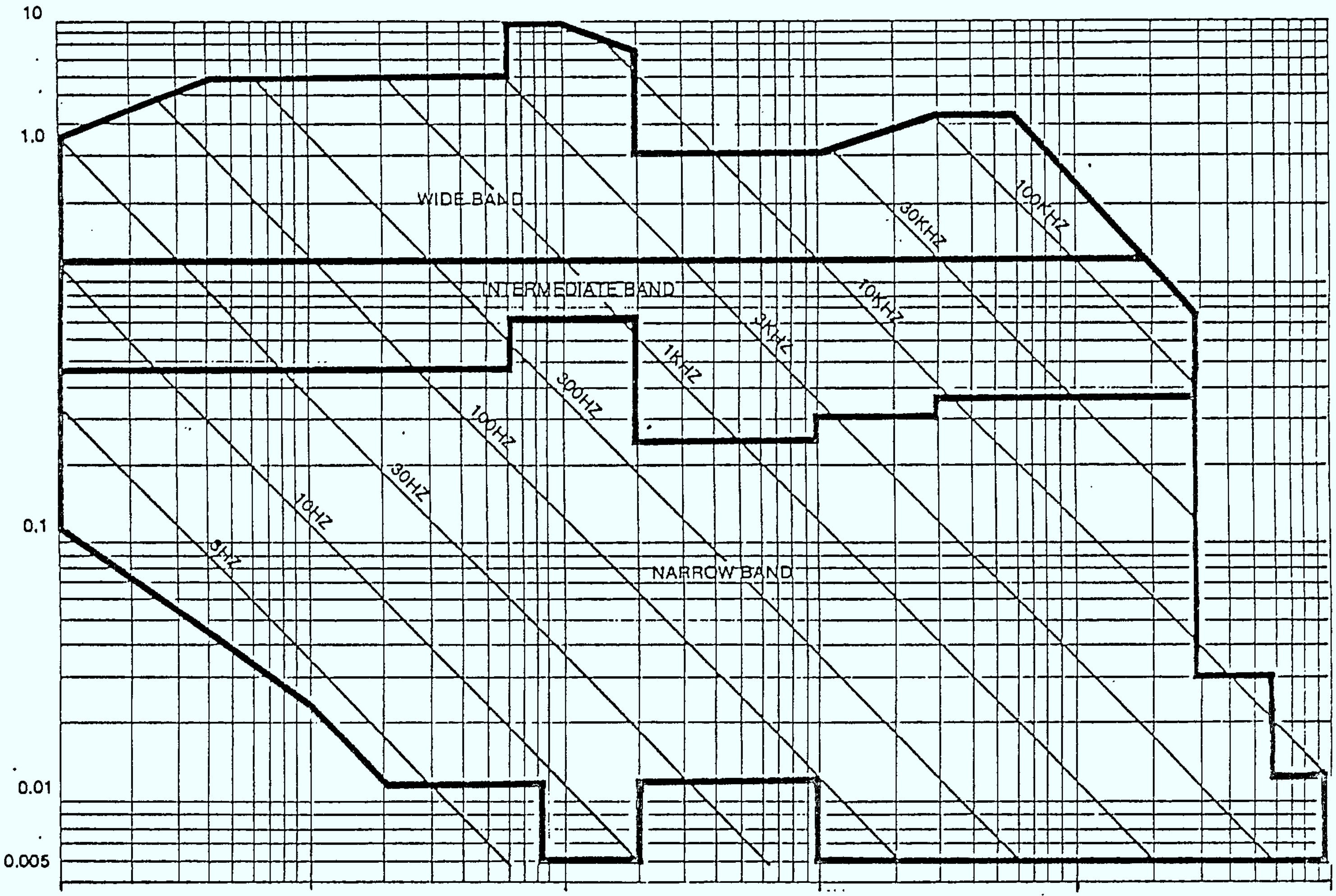
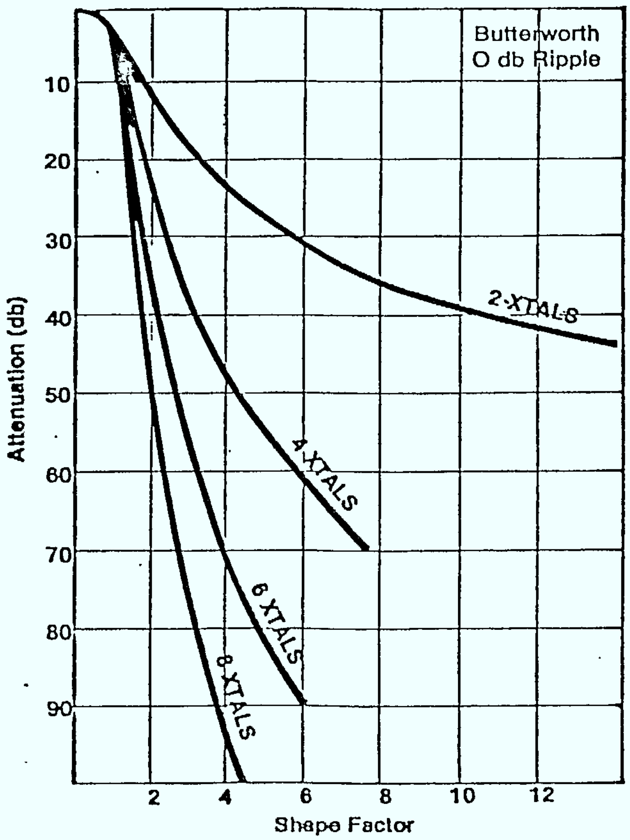


FIGURE A.8 PERCENT BANDWIDTH Vs. CENTER FREQUENCY (ERIE FREQUENCY CONTROL CRYSTAL FILTERS)



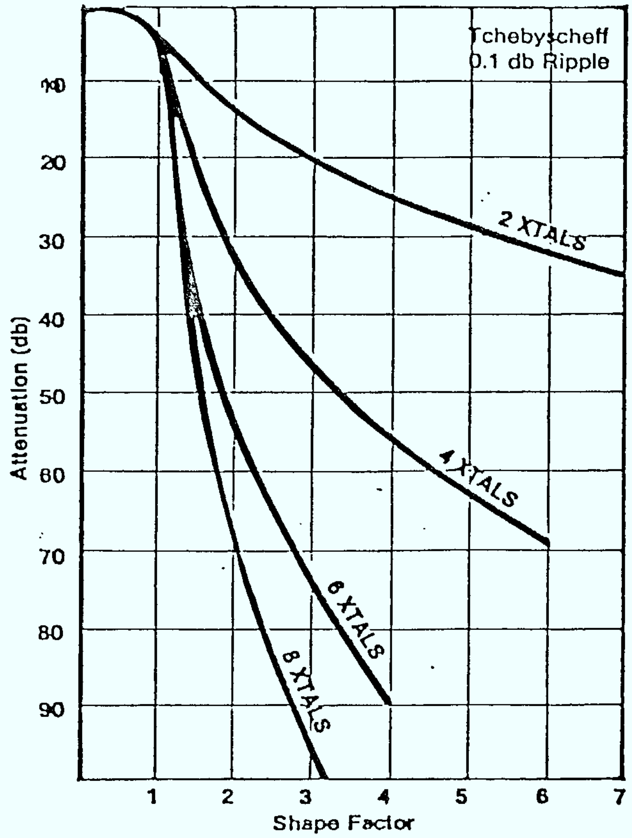


FIGURE A.9 SHAPE FACTOR VS. DESIGN APPROACH FOR TYPICAL CRYSTAL FILTERS (ERIE)

The second method is to utilize two or more crystals in parallel in one or both arms of the filter circuit. This method generally provides less insertion loss and sharper corners, and requires fewer components.

A.3.2 Performance Characteristics

For any given number of sections in a filter, many different shape factors may be achieved depending upon the placement of the frequencies of peak attenuation. The maximum attenuation of frequencies far removed from the passband is obtained when the peak attenuation frequencies are placed at zero and infinity. Moving these frequencies closer to the passband edges gives steeper sides to the filter response curve. However, it also causes the attenuation to decrease or "bounce back" at frequencies further removed from the passband region. The steeper the sides of the response for any given filter, the higher the "bounce back." In the case of single sideband (SSB) filters, where asymmetrical response is desired, all the peak attenuation frequencies are place on the steep side. Of course, the response opposite the steep side then exhibits a rather gentle slope.

Phase shift characteristics of crystal filters are quite closely related to the amplitude characteristics. In general it can be said that the steeper the response curve, the more nonlinear the phase shift becomes, and the steeper the phase shift slope. However, a linear phase response is not nearly as important as the frequency content of a signal since the physiology of the human ear (the final receiver) is such that it is fairly insensitive to relative phase shifts between frequencies.

Figure A.10(a) shows a typical phase shift characteristic of a four-pole filter through the important part of the passband. Neither shape factor nor the phase characteristics are directly dependent upon absolute frequency or bandwidth. Therefore, the frequency axis on the typical curve is labeled in terms of a frequency variable, X, so that the curve is representative of a typical four-pole filter regardless of center frequency or bandwidth. The frequency variable X is defined in the following manner:

$$X = \frac{f-f_o}{B/2} \quad \text{where,}$$

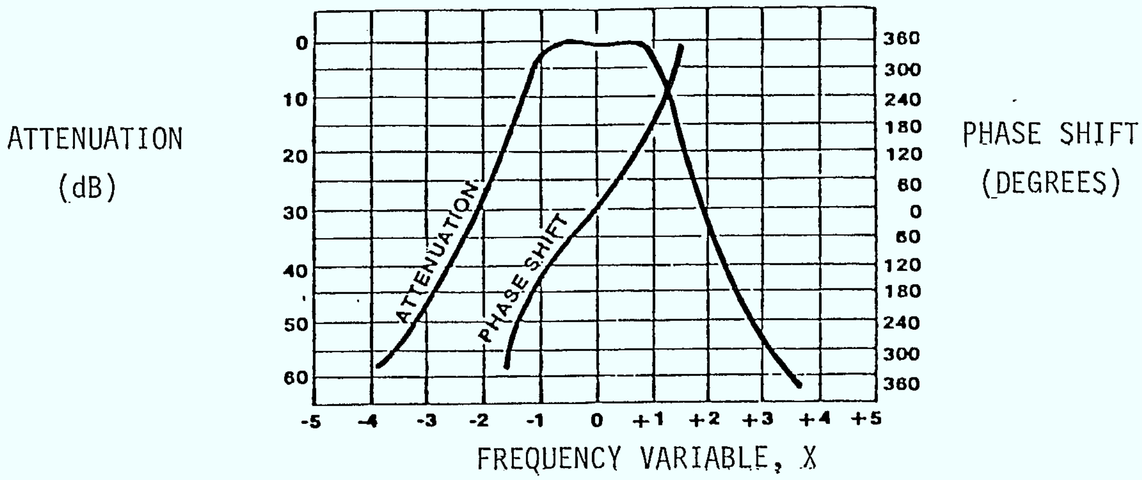
f_o = center frequency of the filter in question

f = frequency of interest

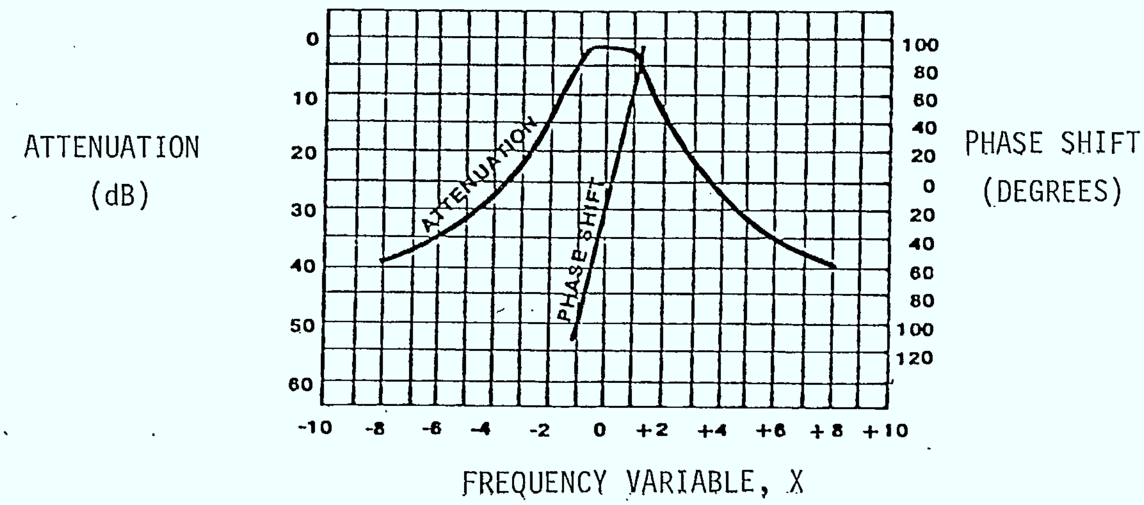
B = bandwidth at 3 dB

Figure A.10(b) shows a typical amplitude/phase response for a two-pole filter.

Input and output impedences vary quite widely throughout the passband. In the case of a narrowband filter, the impedance generally is very low at the low cutoff frequency, and quite high at the high cutoff frequency. For a wideband filter there are two typical impedance characteristics. In the case where series inductance is used, the impedance characteristic is mound-shaped with the highest point at center frequency, and the band edges being the lowest impedance points. When inductance is parallel with the crystals is used, the impedance characteristics is bowlshaped. However, it is sufficient to specify only that certain response



(a) 4-Pole



(b) 2-Pole

FIGURE A.10 TYPICAL AMPLITUDE/PHASE SHIFT CHARACTERISTICS OF A CRYSTAL FILTER: (a) 4-Pole; (b) 2-Pole or phase characteristics are achieved when the filter is driven from a fixed source impedance and terminated in a fixed load impedance. To specify actual input and output impedances of the filter puts severe limitations on the filter design, adds substantially to the cost, and usually serves no useful purpose.

Source and load impedances should be resistive whenever possible. In cases where this is not possible, the reactive component should be specified so that the filter will be aligned under the same conditions as it will be used. The actual values of the required source and load impedances are not particularly critical to the initial filter design, since it is a simple matter to transform the natural filter impedance to that required. In most cases, these transformers are integral parts of the filter design. Once established, however, these impedances should be within $\pm$ 5% of that specified. Consideration must be given to the signal generator impedance, and any stray impedances, such as meter or cable capacity, when checking for proper termination.

While it is difficult to generalize on cost of filters due to the tremendous number of variables, it is usually found that Intermediate Band filters are most expensive, followed by Wide Band, and the least expensive are Narrow Band.

A.3.3 Suitability For AM Receiver Applications

Although it is possible to obtain crystal filters at 455 kHz with 3 dB bandwidths in the range of 6 to 9 kHz, the number of inductances required to achieve this wide bandwidth would make the size of the filter too bulky* to be competitive with either IF transformers or ceramic filters. Furthermore, because of the rather wide bandwidth required for the crystal filter, it is needless to say, not a standard product line of crystal filter manufacturers. No information as to the possible cost of such a filter was obtained but indications are that it would be fairly expensive in view of the custom design requirement.

It seems fairly safe to conclude, therefore, that crystal filters are not really viable for AM receiver applications.

A.4 Mechanical Filters

Although the commercial use of mechanical filters is far more recent than that of crystal filters, their basic operating principles are well established. The mechanical filter is a mechanically resonant device which receives electrical energy, converts it into mechanical vibration, and then converts the mechanical energy back into electrical oscillations at the output. The reason for this double conversion is simply that the mechanically resonant system is better at suppressing unwanted frequencies than a completely electrical system of the same size. This, in turn, is due to the fact that mechanical (i.e., acoustic) vibrations have a much lower velocity than electrical oscillations, and hence the mechanical wavelengths are much shorter. It is thus possible to use mechanical resonators having half-wavelength dimensions, and obtain the same sort of high

* As an indication, one particular 455 kHz crystal filter with a 3 dB bandwidth of only 2 kHz and a shape factor (60 dB: 3 dB) of 2.8 measures 3.2 "(L) x 1.3"(W) x 1.1"(H).

Q as that obtainable with resonant transmission lines.

A.4.1 Operating Principles

The mechanical filter, as illustrated in Figure A.11(a) consists of an input electromechanical transducer, an output mechanoelectrical transducer operating like the input one but in the reverse direction, mechanical resonators in the form of metal discs, and wire couplers to couple the resonators to each other. The equivalent circuit of Figure A.11 (b) is the electrical analog of the mechanical filter. It shows that the resonant discs correspond to shunt parallel-tuned circuits L_1C_1, the mechanical coupling wires are analogous to the series inductances L_2, and coil losses are shown as series lumped resistors, and contribute to both filter impedance and transmission losses.

After the input signal has been converted into mechanical vibrations, these are applied to the first section of coupling wires, which then vibrate longitudinally and apply torsional oscillations to the first disc. These oscillations are then transmitted to the remaining sections of the filter, and the output is converted back into electrical oscillations. As the electrical equivalents of the mechanical elements are very high-Q circuits, the bandwidth can be made as narrow as desired, with an extremely sharp attenuation outside the passband.

The transducer, which converts electrical energy into mechanical vibrations or vice versa, may be either a magnetostrictive device or an electrostrictive one, the former being decidedly the more popular of the two. The magnetostrictive transducer is based on the principle

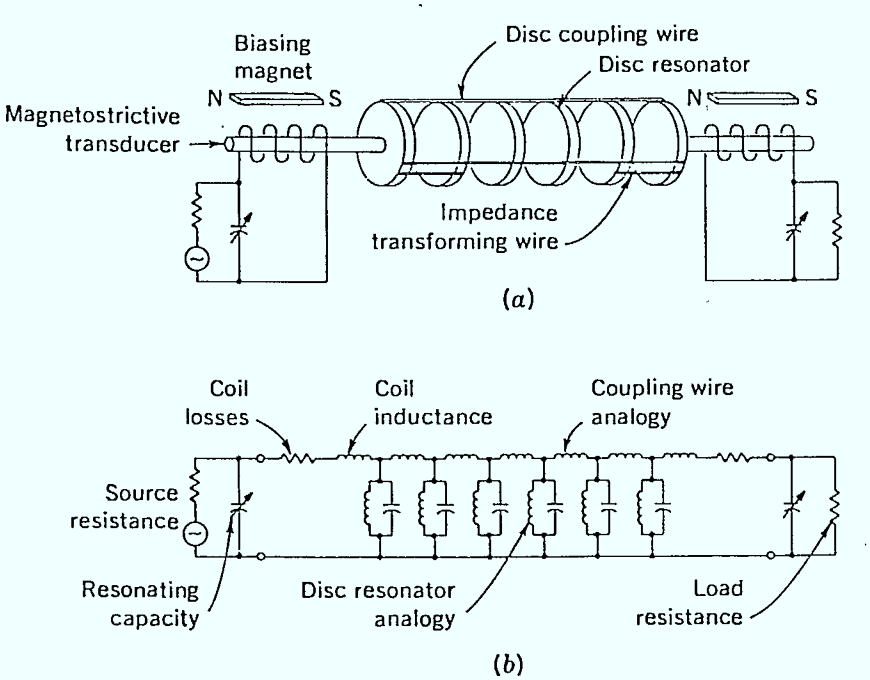


FIGURE A.11 THE MECHANICAL FILTER* (a) PHYSICAL CONSTRUCTION: (b) ELECTRICAL ANALOGY

*This particular illustration is of the mechanical filter manufactured by Rockwell-Collins Divisions.

that certain materials, notably nickel and ferrites, elongate or shorten in the presence of a magnetic field. Thus, if a current is sent through a coil whose core is magnetostrictive, mechanical oscillations result and can be used to drive the mechanically resonant elements of the filter. The electrostrictive transducer, on the other hand, is based on the fact that some materials, such a piezoelectric crystals and the ceramic barium titanate, will compress and expand in an electric field; this type is thus voltage-activated. The transducer not only converts energy into the appropriate form but also contributed to the terminations of the filter; both of these functions are considered in its design.

A.4.2 Filter Characteristics

The equivalent circuit of Figure A.11(b) shows that the center frequency of the mechanical filter is determined by the resonators, as represented by shunt L_1C_1 parallel-tuned circuits; taking into account physical dimensions, permissible tolerances, and manufacturing and tuning problems, filters with center frequencies from about 60 to 500 kHz are being manufactured. Since each resonator represents a tuned circuit, it follows that an increase in the number of discs will improve the selectivity, or shape factor, of the mechanical filter.

The coupling inductors L_2 of the equivalent circuit represent the coupling wires; by varying this mechanical coupling, it is possible to adjust the bandwidth of the filter accordingly. Because the bandwidth varies roughly as the total area of the coupling wires, it may be increased either by using thicker coupling elements or by using more of them. These considerations result in filters with bandwidths in the 0.5-to-50 kHz range being commercially available.

The mechanical-filter frequency response of Figure A.12, shows the attenuation provided by a typical single sideband (SSB) mechanical filter and may be used to explain the various terms used in connection with mechanical filters. Thus, it is seen that the 6 dB bandwidth is 3.5 kHz and the 60 dB bandwidth is 4.2 kHz, yielding a shape factor of 1.2, which is typical of this kind of filter. Although an ideal filter should have a flat "nose," or passband, practical limitations prevent this ideal from being obtained with mechanical filters (or any other sort, for that matter). The term *ripple amplitude*, or *peak-to-valley ratio*, is used to specify this passband characteristics and is the ratio of maximum to minimum attenuation within the filter's useful passband. The value shown in Figure A.12, just under 3 dB, is common, but peak-to-valley ratios in the vicinity of 1 dB are also available.

Spurious responses occur in mechanical filters because of resonances other than the desired ones,* but (unlike the responses in crystal filters) these may easily be kept far enough from the passband to permit other tuned circuits to attenuate the spurious passbands that may be produced, although these spurious passbands can be made so far down on the main response that external tuned circuits are no longer necessary.

*For example, flexure unstead of longitudinal vibration for the couplers, and/or bending or shear vibration instead of tosion for the discs.

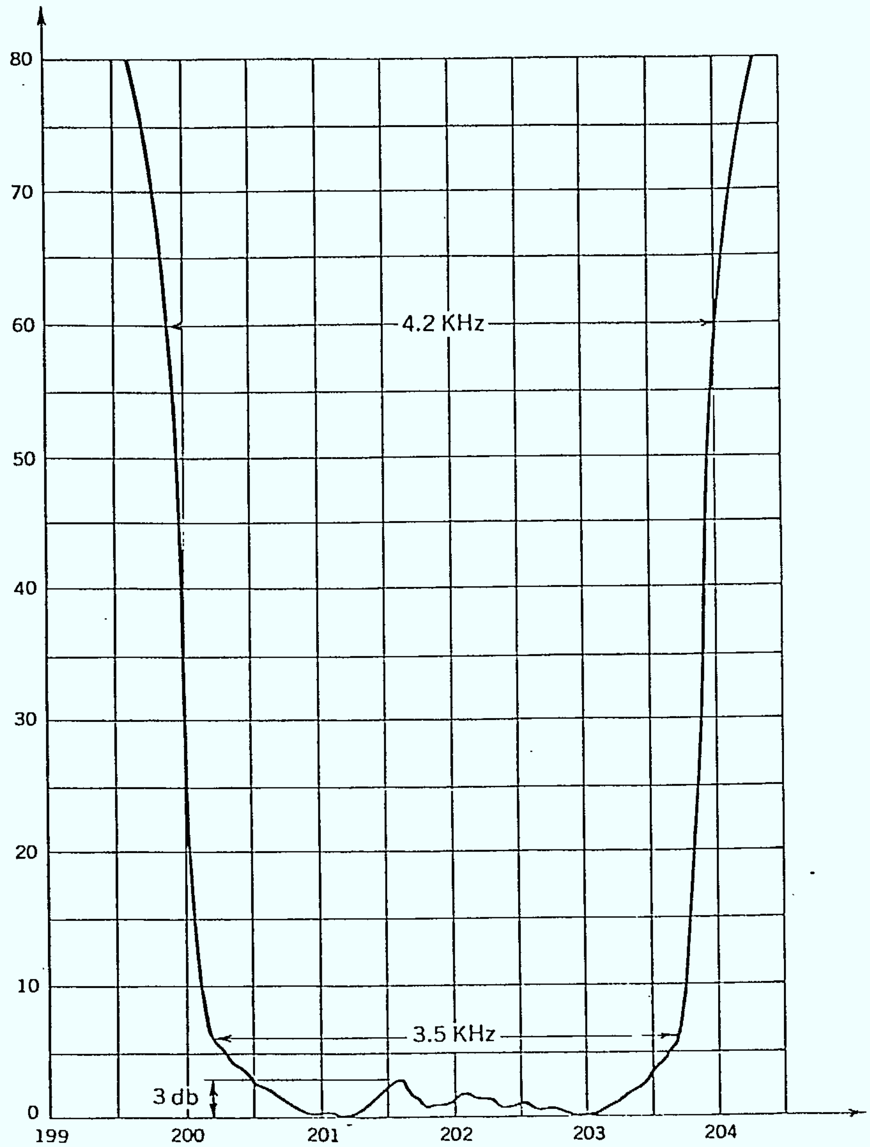


FIGURE A.12 TYPICAL SINGLE SIDEBAND FREQUENCY RESPONSE FOR A MECHANICAL FILTER

A.4.3 Suitability For AM Receiver Applications

Small mechanical filters operating at 455 kHz are readily available from Rockwell-Collins Division of California with a variety of passbands ranging from 0.2 kHz to 16 kHz. One particular filter exhibits a 3 dB passband of 8 kHz and a 60 dB bandwidth of 24 kHz, i.e., a shape factor (60 dB: 3 dB) of 3:1. The case dimensions for this filter are typically 2.74"(L) x 0.5" (W) x 0.62" (H).

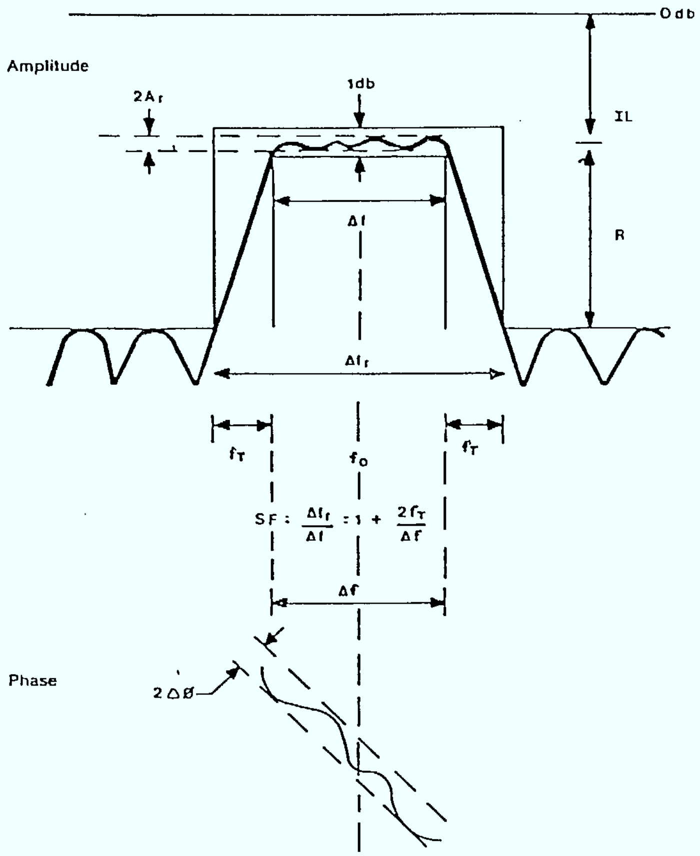
While the mechanical filter is undoubtly capable of providing very high selectivity in AM receivers, particularly when two or more are connected in tandem, its fairly large size and cost ($60.00 for quantity 500) do not make it particularly attractive for application in any AM receiver.

A.5 Surface Acoustic Wave (SAW) Filters

The Surface Acoustic Wave (SAW) bandpass filter is a miniature device which exhibits flat amplitude response and linear phase and has been used in a wide variety of IF applications (notably in domestic television receivers) Representative amplitude and phase characteristics for a symmetrical SAW filter are illustrated in Figure A.13 together with a definition of the key descriptive parameters.

A.5.1 Design Characteristics And Performance

SAW filters are implemented by constructing the filter's impulse response on the surface of piezoelectric substrate



- f_o Center frequency
- Δf Bandwidth typically defined at the —ldb point
- Δf_r Bandwidth at the ultimate rejection level
- f_T Transition width typically measured from band edge to the ultimate rejection level
- SF Shape factor defined as the ratio of Δf_r to Δf
- A_r Peak amplitude ripple in Δf.
- IL Insertion loss at f_o
- R Ultimate rejection level referenced to f_o
- $\Delta\phi$ Peak phase deviation from linearity in Δf.

FIGURE A.13 SPECIFICATION PARAMETERS FOR SAW IF BANDPASS FILTERS.

using a transversal filter tapped delay-line approach. The fabrication technique uses planar processing methods ensuring reproducibility and low cost. The bandpass response determined by the transducer design, is permanently established by a single master pattern so that tuning adjustments are never required.

The two most common SAW substrate materials are ST Quartz and YZ lithium niobate. ST Quartz has an extremely low temperature coefficient (.03 ppm/$(^{o}C)^2$ referenced to $25^{o}C$ and is useful for fractional bandwidths of up to 5% of f_o. Lithium niobate is used for the wider bandwidth requirements since it offers a lower insertion loss. Its temperature coefficient is 90ppm/^{o}C. The range of performance parameters for the two methods is indicated in Figure A.14. The temperature coefficient applies to the center frequency of the response. Relative bandwidth remains constant with the temperature.

The key parameter in the design of SAW filters is the transition width f_T. The length of the substrate material and the complexity of the design is inversely proportional to f_T. Once f_T is specified the filter can be designated for any Δf in the range from 100 kHz to 40% of f_o. A practical lower limit on f_T is 300 kHz. The achievable shape factor is then a function of Δf. This relationship is plotted in Figure A.14 for three values of f_T. The horizontal lines of Figure A.14 show the achievable shape factors (50 dB: 3 dB) for several conventional Chebyshev and Butterworth LC designs. It is clear that for bandwidths of greater than a few hundred kiloHertz, SAW filters offer lower shape factors and superior selectivity with characteristics approaching that of the ideal rectangular bandpass filter.

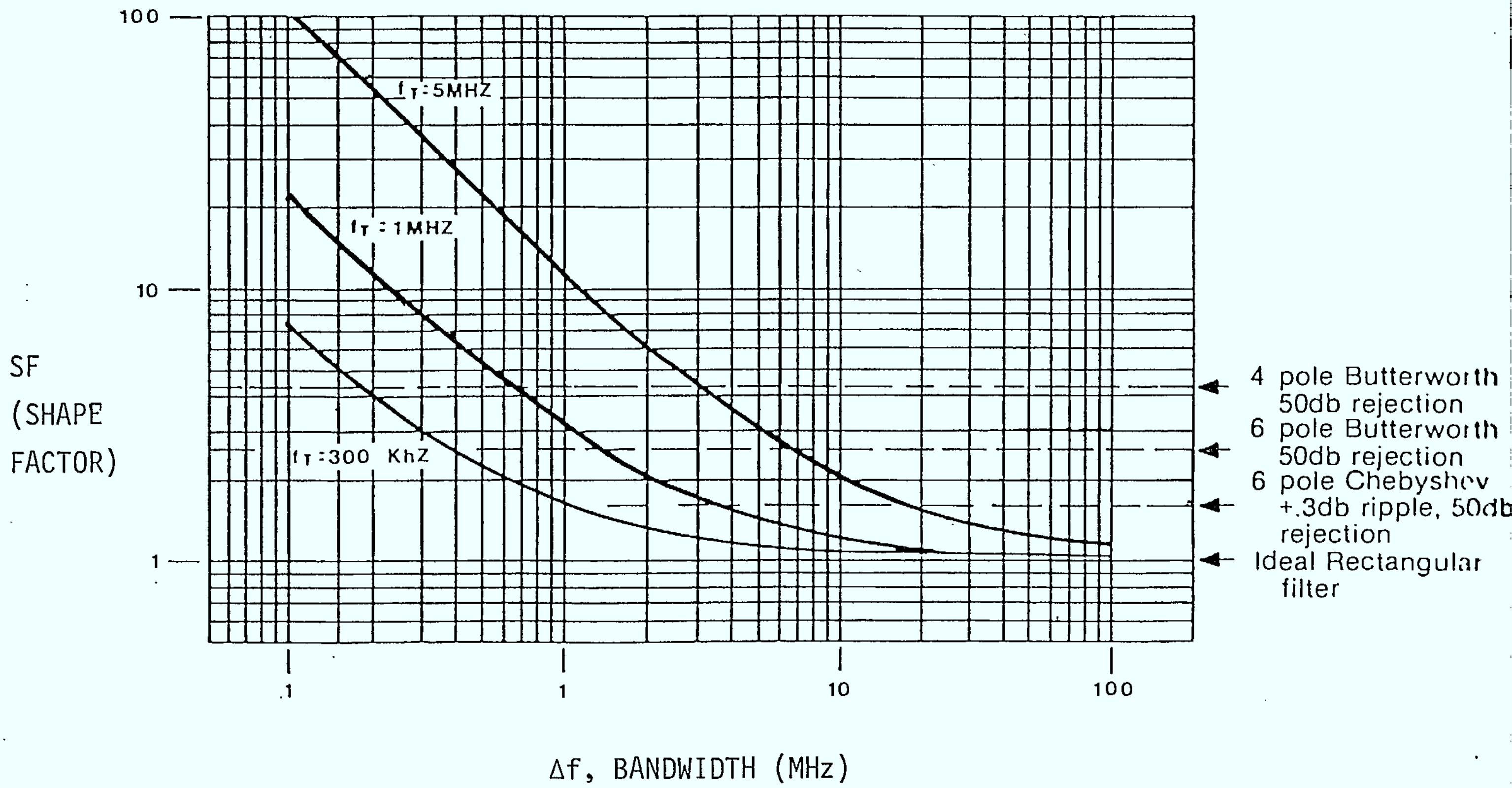


FIGURE A.14 SHAPE FACTORS FOR SAW IF FILTERS

The total amplitude ripple in a SAW filter ranges from ± 2 dB to ± 6 dB depending on the design.

SAW filters uniquely provide performance in the shaded area of Figure A.15 covering 1 dB bandwidths from 100 kHz to 400 kHz at center frequencies above 40 MHz. Crystal filters must operate on harmonics and are too narrowband while LC and cavity filters do not have sufficient Q and are too broadband. Two SAW units can be cascaded to realize over 80 dB of out-of-band rejection.

The ultimate rejection of SAW filters is in the 30 to 60 dB range depending on design complexity. Most filters falling within the solid lines of Figure A.15 can be designed for an ultimate rejection in the 40 - 50 dB range.

SAW filters differ from conventional filters in that they exhibit a time delay in the 1 to 5 μsecond range. In most cases offensive time echoes are attenuated 40 - 45 dB below the desired output level.

SAW filters typically have insertion losses in the 15 to 30 dB range when operating with a source and load impedance of 50 ohms. In general, insertion loss increases with increasing bandwidth, increasing ultimate rejection and decreasing amplitude ripple.

The size of a SAW filter is dependent on transition width. For transition widths ranging from 300 kHz

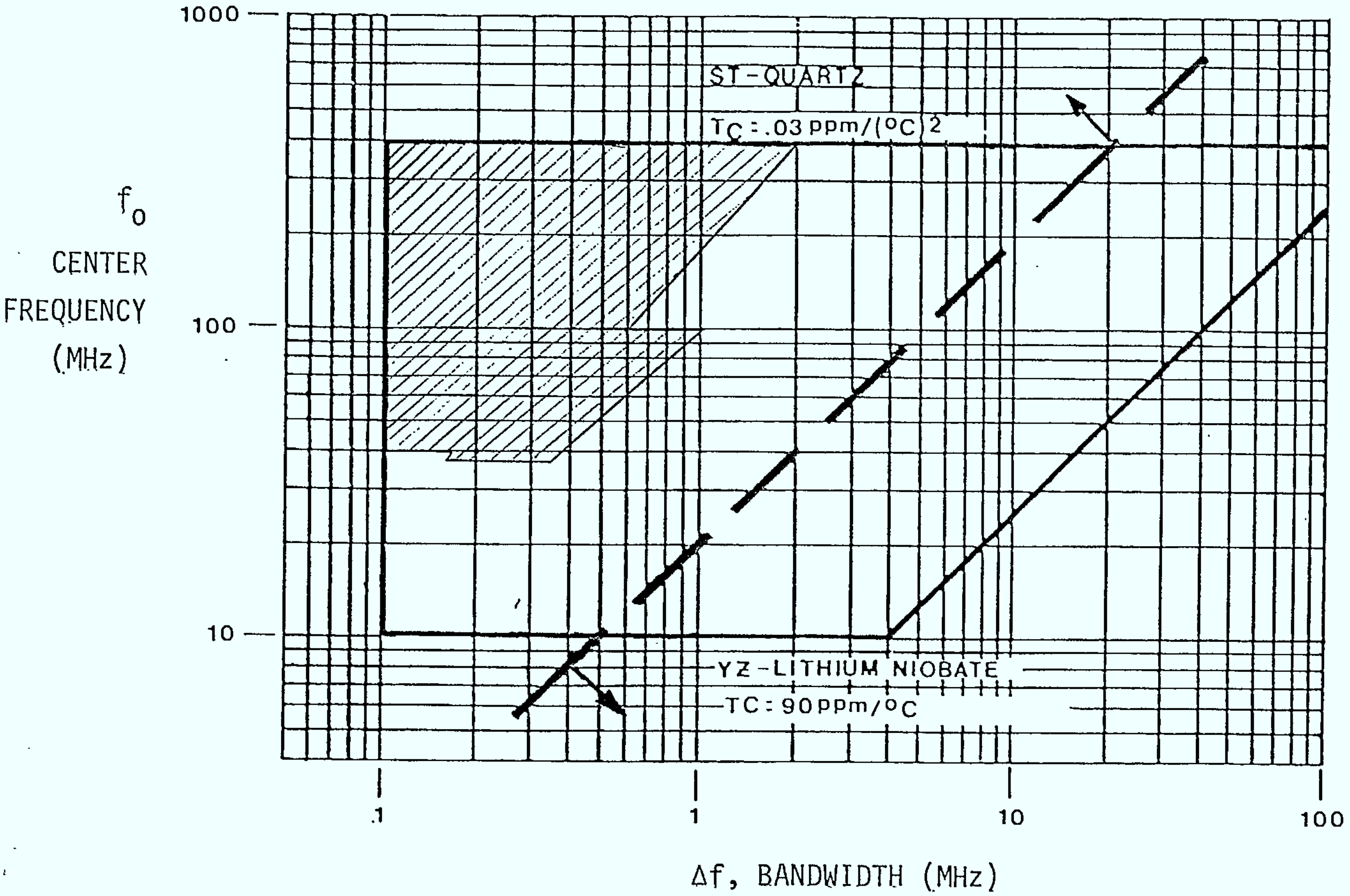


FIGURE A.15 BANDWIDTH, CENTER FREQUENCY AND TEMPERATURE COEFFICIENT FOR SAW IF FILTERS to 5 MHz the crystal size will decrease in length from near 2" to a fraction of an inch. Typical packages range from a TO-8 can for the smaller sizes to a pin mounted flat pack for the larger lower transition width units.

A.5.2 Suitability For Application in AM Receivers

From the preceding discussion, it is obvious that SAW filters cannot be used for IF applications in AM receivers because of the lower limit on band-pass center frequency. This lower limit is set by practical size limits on substrates and the range of velocities of available surface wave materials. At center frequencies lower than 10 MHz, the SAW bandpass filter becomes unattractively bulky and becomes less competitive with other conventional filter technologies (e.g., bulk crystal, LC).

A.6 Ceramic Filters

A ceramic filter is an electromechanical filter employing a ceramic piezoelectric material and having bandpass characteristics at either 455 kHz or 10.7 MHz, the two most common IF frequencies used in receivers.

A.6.1 Theory of Operation

All ceramic filters derive their basic frequency selective capability from a mechanical vibration that results from a piezoelectric effect in the ceramic material. While the theoretical analysis of this piezoelectric effect is a relatively complex electromechanical function, it can be shown as a simple equivalent circuit as illustrated in Figure A.16.

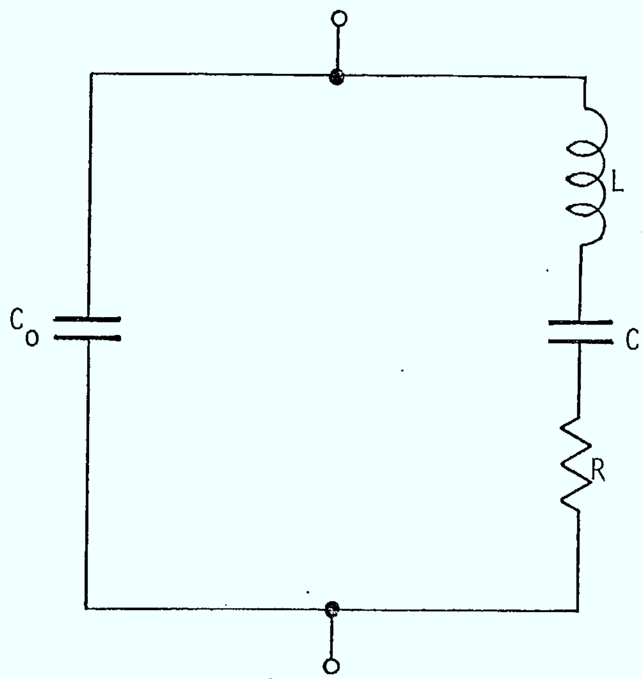


FIGURE A.16 EQUIVALENT CIRCUIT FOR THE PIEZOELECTRIC EFFECT IN CERAMIC MATERIALS

This equivalent circuit represents a typical 2-terminal device, which is the basic building block for more complex filters, where C_o is equal to the static capacity of an equivalent ceramic capacitor, while L, C and R represent equivalent mass, equivalent compliance and equivalent resistance respectively.

The resonant frequency, for this 2-terminal filter element can be calculated as follows:

$$fo = \frac{1}{2\pi\sqrt{LC}}$$

and antiresonance as:

$$f_\infty = \frac{1}{2\pi\sqrt{\frac{CC_o}{C + C_o}}}$$

In addition to the basic 2-terminal configuration, it is possible to create a 3-terminal filter as shown in Figure A.17. This filter configuration, through proper design of the element, can provide different input and output impedances, a characteristic that is particularly useful for interstage coupling between semiconductor amplifiers.

When 2 or more basic 2-terminal elements are connected in a 4-terminal arrangement, as shown in Figure A.18, substantial increases in selectivity can be obtained, dependent upon the number of elements utilized. These 4-terminal arrangements are extremely effective as coupling devices between IF amplifier stages since thay are easily impedance matched, have low insertion loss, and require no tuning while increasing selectivity far beyond that possible with standard coupling transformers. Additionally, by controlling the coefficient of electromechanical

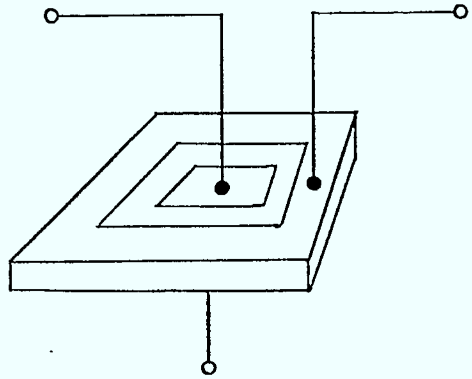


FIGURE A.17 3-TERMINAL CERAMIC FILTER

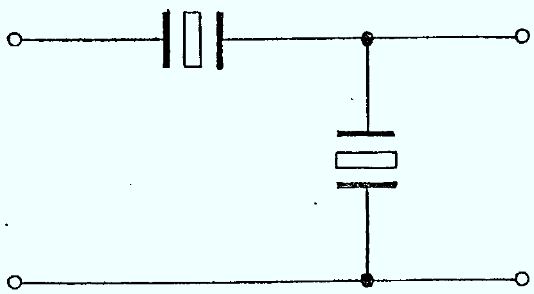


FIGURE A.18 4-TERMINAL CERAMIC FILTER

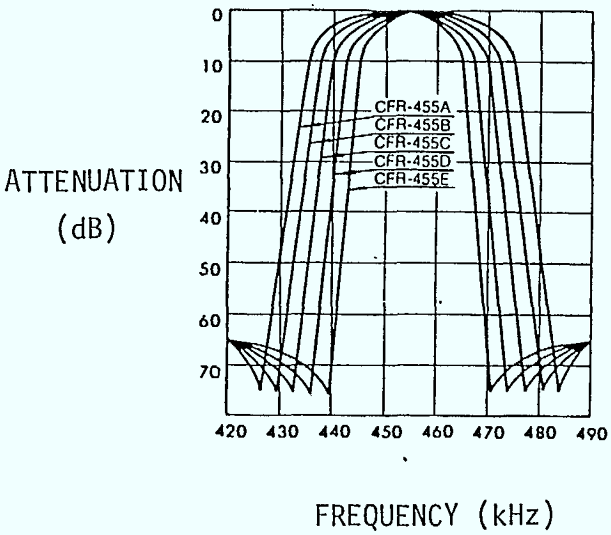
coupling between filter elements (K), bandwidth can be accurately controlled to provide either a very "peaked" response characteristic or a "flat-top" response characteristic as required by 10.7 MHz FM and 4.5 MHz TV IF systems. A typical set of 455 KHz response curves for ceramic filters manufactured by Murata Corporation is shown in Figure A.19.

A.6.2 Suitability For AM Receiver Applications

The Type CFR-455 filter of Figure A.19 is of ladder-type construction utilizing 11 ceramic resonators and is intended primarily for use in communications receivers. Table A.2 lists the performance specifications for these filters while Figure A.20 indicates the dimensions. Note, for example, the excellent selectivity at ± 10 kHz (adjacent channel attenuation - ACA) provided by the CFR-455G.

As expected, this particular series of ceramic filters is fairly expensive at approximately $10.50 each (in quantity 500) but might well be considered for use in the premium-priced AM/FM stereo console receivers and especially in automotive receivers which are often subjected to a high interference environment.

For a more universal application to all types of AM receivers with the exception of possibly the portable pocket AM radio, one would be more inclined to consider the lower cost and hence lower performance filters such as the Type CFW-455 and CFU-455 filters, also produced by Murata Corporation. Tables A.3 and A.4 summarize the performance of these two types of filters which, like the CFR-455, are also intended for use in communications receivers. Note however, that a single filter (eg. CFW-455H or CFU-455H2) still provides an adjacent channel selectivity that far exceeds that of a



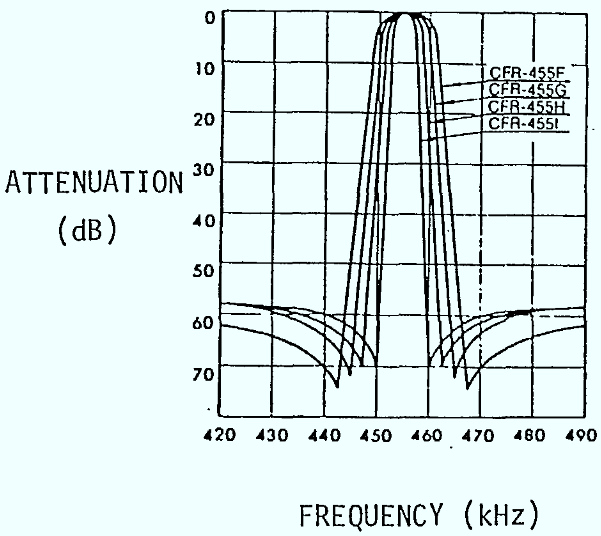


FIGURE A.19 CERAMIC 455 kHz IF BANDPASS FILTER RESPONSE CURVES* (FOR COMMUNICATIONS RECEIVERS)

*Murata Corporation of America

MODEL	Center Frequency (KHz)	3dB Band Width (KHz) Min.	6dB Band Width (KHz) Min.	60dB Band Width (KHz) Max.	Spurious (dB) Min.	Insertion Loss (dB) Max.	In, Output Impedance (Ω)	Temperature Range (°C)
CFR-455 A	455	±13	±17.5	±30	45	6	1000	−20 to +80
CFR-455 B	455	±10	±15	±25	45	6	1000	
CFR-455 C	455	± 9	±13	±23	45	6	1000	
CFR-455 D	455	± 7	±10	±20	45	6	1500	
CFR-455 E	455	± 5.5	± 8	±16	40	8	1500	
CFR-455 F	455	± 4.2	± 6	±12	40	8	2000	
CFR-455 G	455	—	± 4	±10	40	8	2000	
CFR-455 H	455	—	± 3	± 7.5	40	9	2000	
CFR-455 I	455	—	± 2	± 5	40	10	2000	

Stability of Center Frequency: Within 0.4% for 10 years Within 0.3% from −20°C to +80°

TABLE A.2 TYPE CFR-455 CERAMIC FILTER PERFORMANCE SPECIFICATIONS

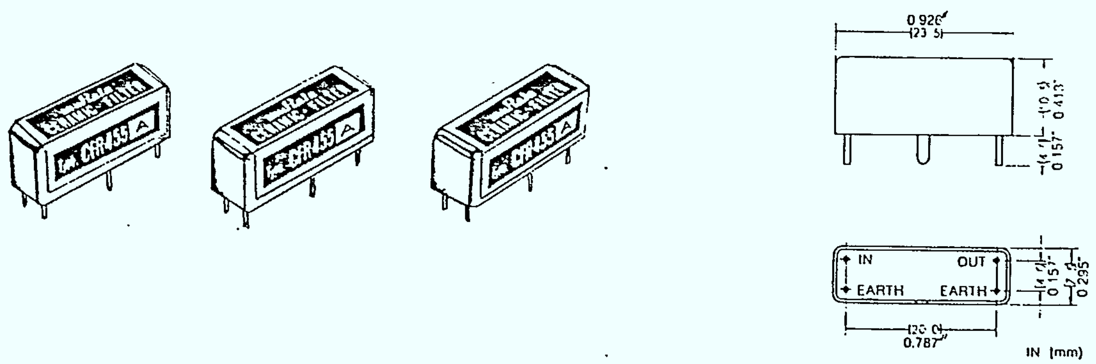


FIGURE A.20 TYPE CFR-455 FILTER DIMENSIONS

MODEL	Center Frequency (KHz)	6dB Band Width (KHz) min.	50dB Band Width (KHz) Max.	Spurious Response (dB) Min.	Insertion Loss (dB) Max.	In, Output Impedance (Ω)
CFW455B	455	±15	±30	35	4	1500
CFW455C	455	±12.5	±24	35	4	1500
CFW455D	455	±10	±20	35	4	1500
CFW455E	455	± 7.5	±15	35	6	1500
CFW455F	455	± 6	±12.5	35	6	2000
CFW455G	455	± 4.5	±10	35	6	2000
CFW455H	455	± 3	± 9	35	6	2000
CFW455I	455	± 2	± 7.5	35	7	2000

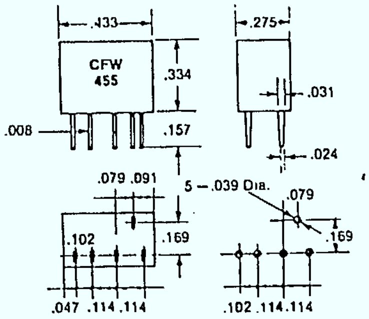


TABLE A.3 TYPE CFW-455 FILTER PERFORMANCE SPECIFICATIONS

MODEL	Center Frequency (KHz)	6dB Band Width (KHz) min.	40dB Band Width (KHz) Max.	Spurious Response (dB) Min.	Insertion Loss (dB) Max.	In, Output Impedance (Ω)
CFU455B2	455±2	±15	±30	27	4	1500
CFU455C2	455±2	±12.5	±24	27	4	1500
CFU455D2	455±1.5	±10	±20	27	4	1500
CFU455E2	455±1.5	± 7.5	±15	27	6	1500
CFU455F2	455±1.5	± 6	±12.5	27	6	2000
CFU455G2	455±1	± 4.5	±10	25	6	2000
CFU455H2	455±1	± 3	± 9	25	6	2000
CFU455I 2	455±1	± 2	± 7.5	25	6	2000

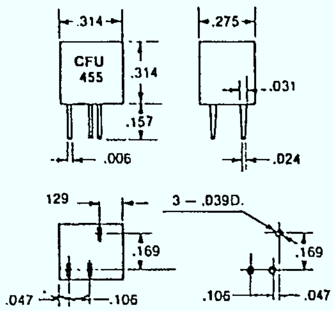


TABLE A.4 TYPE CFU-455 CERAMIC FILTER PERFORMANCE SPECIFICATIONS conventional 2-stage IF amplifier using three single-tuned IF transformers. The CFW-455 is slightly preferable to the CFU-455 because of its better selectivity and spurious response characteristics.

The approximate costs for the Type CFW-455 and Type CFU-455 filters are \$5.15 and \$3.50 respectively for quantity 500 which compare very favourably with the cost of a single-tuned IF transformer (\$2.61 in quantity 500).

The small size, relatively low cost and high selectivity of ceramic filters make them strong contenders for replacing IF transformers in AM receivers. Another key advantage is that since ceramic filters require no tuning or adjustment, they would obviate the need for aligning or tuning the IF stages which could amount to a significant saving in labour costs and would facilitate possible automated assembly.