TRANSMITTERS FOR FM BROADCASTING

THE FM BROADCAST TRANSMITTER

The purpose of the FM transmitter is to convert one or more audio frequency (composite baseband) input signal or an AES3 serial digital audio data bit stream into a frequency-modulated, radio-frequency (RF) signal at the desired power output level to feed into the radiating antenna system. In its simplest form, the FM broadcast transmitter can be considered an FM modulator and an RF power amplifier packaged in one unit.

Actually the FM transmitter consists of a series of individual subsystems each having a specific function:

1. The FM exciter converts the analog audio baseband or serial digital audio data into frequency-modulated RF and determines the key qualities of the signal.
2. The intermediate power amplifier (IPA) is required in some transmitters to boost the RF power level up to a level sufficient to drive the final stage.
3. The final power amplifier further increases the signal level to the final value required to drive the antenna system.
4. The power supplies convert the input power from the ac line into the various dc or ac voltages and currents needed by each of these subsystems.
5. The transmitter control system monitors, protects, and provides commands to each of these subsystems so that they work together to provide the desired result.
6. The RF low-pass filter removes undesired harmonic frequencies from the transmitter's output, leaving only the fundamental output frequency.
7. The directional coupler provides an indication of the power being delivered to and reflected from the antenna system.

Figure 1 shows a simplified block diagram of a typical FM transmitter.

FM BROADCAST TRANSMITTER POWER OUTPUT REQUIREMENTS

The FCC regulates the power of FM broadcast stations in terms of effective radiated power (ERP) which is determined by the class of station and the antenna height above the average terrain. The authorized ERP applies only to the horizontally polarized component of radiation. Elliptical or circular polarization is also permitted where the ERP of the vertically polarized component may be as great as the authorized horizontal component. This means that twice as much total power is radiated and twice as much transmitter power is required.

The transmitter power requirement is reduced by increasing the gain of the antenna. There is, of course, an economic tradeoff between the cost of a higher gain antenna versus the cost of a larger transmitter and the added primary power costs. For a high ERP, it is common to use antennas with up to 12 elements which provide a power gain of about 12.6 (or 6.3 in each polarization).

The long transmission lines associated with the tall towers commonly used are a source of considerable power loss. For example, the efficiency of 2000 ft of $3\frac{1}{8}$ in. rigid coax at 100 MHz is only about 62%.

The required ERP is first determined. Then the transmitter power output (TPO) can be calculated taking into account the transmission line losses and the antenna array gain. Depending on the particular situation, the TPO could vary from as little as 50 W to as much as 70 kW.

FM transmitters are designed to operate over a wide range of power outputs so that any required power output can be furnished with a few basic sizes. Popular maximum ratings range from 250 W to 70 kW. Most installations use a maximum TPO of 30 kW or less because it is more economical to achieve the maximum 100 kW of ERP with circular polarization by sufficient antenna gain.

FM Exciters

The heart of an FM broadcast transmitter is its exciter. The function of the exciter is to generate and modulate the carrier wave with one or more inputs (mono, stereo, SCA) in accordance with the FCC standards. Then the FM modulated carrier is amplified by a wideband amplifier to the level required by the transmitter's following stage.

Stereo transmission places the most stringent performance requirements on the exciter. Because the exciter is the origin of the transmitter's signal, it determines most of the signal's technical characteristics, including signal-to-noise-ratio (SNR), distortion, amplitude response, phase response, and frequency stability. Waveform linearity, amplitude bandwidth, and phase linearity must be maintained within acceptable limits throughout the analog baseband chain from the stereo and subcarrier generators to the analog FM exciter's modulated oscillator (1). Figure 2 shows the frequency spectrum of the modulating baseband signal including: monaural, stereo pilot, stereo subchannel, and one SCA subcarrier. The recent introduction of AES3 (Audio Engineering Society Digital Audio Transport Standard) digital audio transport and all-digital FM modulation techniques like direct digital synthesis

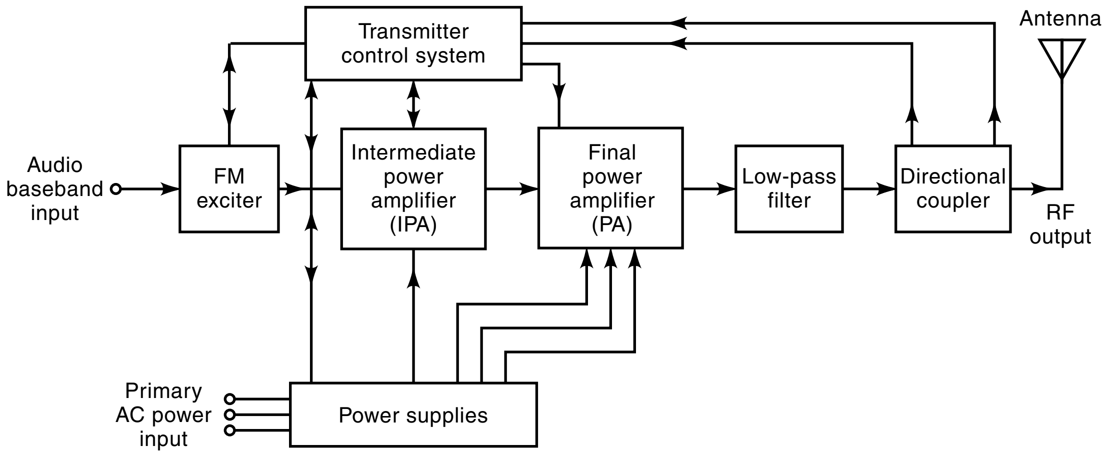


Figure 1. Simplified block diagram of an FM broadcast transmitter.

(DDS) eliminate the distortions introduced by analog circuits. In a digital FM exciter, the left and right audio data are converted into a digital representation of stereo baseband by digital signal processing (DSP). Then these data are further converted into a frequency-modulated carrier by a DDS numerically controlled oscillator (NCO). From here, the FM carrier is usually amplified in a series of class *C* nonlinear power amplifiers, where any amplitude variation is removed. The amplitude and phase responses of all the RF networks that follow the exciter must also be controlled to minimize degradation of the signal quality.

Direct FM

Direct FM is a modulation technique where the frequency of an oscillator can be made to change in proportion to an applied voltage. Such an oscillator, called a voltage tuned oscillator (VTO), was made possible by the development of varactor tuning diodes which change capacitance as their reverse bias voltage is varied (also known as a voltage controlled oscillator or VCO).

If the composite baseband signal is applied to the tuning terminal of a VTO, the result is a *direct FM* modulated oscillator. Figure 3 is a block diagram that describes most of the modern direct FM exciters on the market.

Automatic Frequency Control

The frequency stability of direct FM oscillators is not good enough to meet the FCC frequency tolerance of ±2,000 Hz. This requires an automatic frequency control system (AFC) that uses a stable crystal oscillator as the reference frequency.

The modulated oscillator need not have good long-term stability because the AFC feedback loop will correct for long-term drift to keep the average carrier frequency within limits. The modulated oscillator does need excellent short-term stability (less than 1 s) because the control-loop time constant must be long enough so that the AFC circuit does not try to remove desired low-frequency audio modulation. This means that the oscillator is essentially running open-loop at frequencies above a few hertz so that the noise performance of the modulator will also be determined by the short-term stability characteristics of the oscillator.

Phase-Locked-Loop Automatic Frequency Control

Phase-locked-loop (PLL) technology has provided a means of precisely controlling the carrier's average frequency while

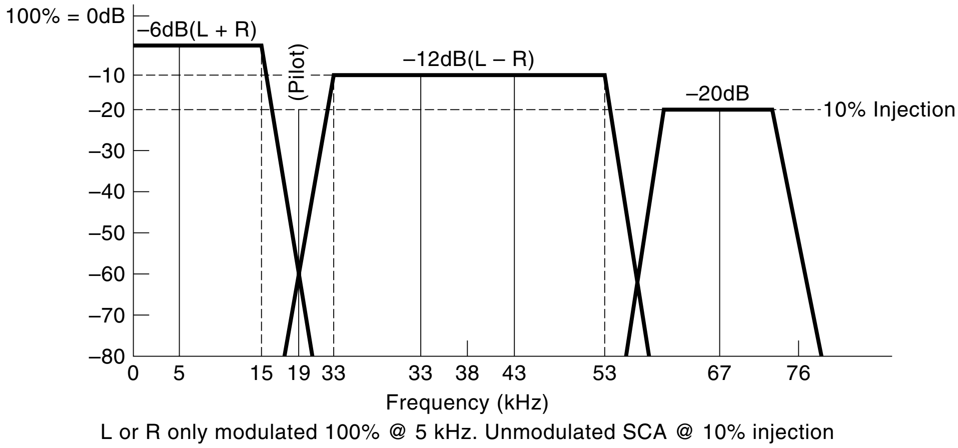


Figure 2. Stereo composite baseband with SCA subcarrier.

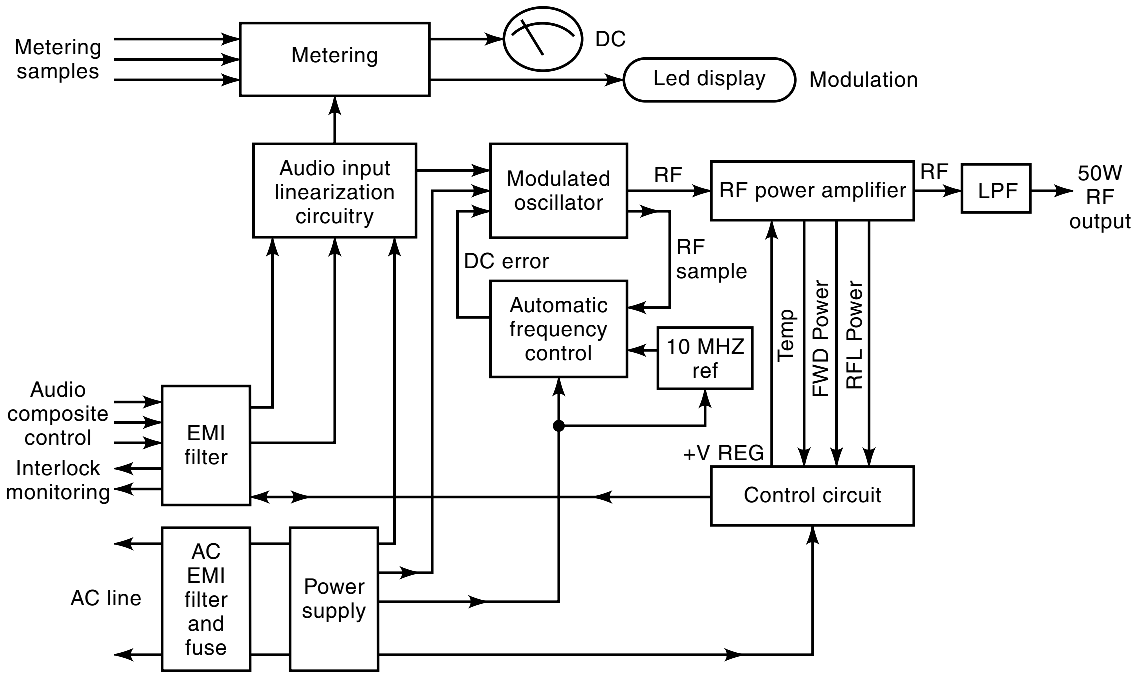


Figure 3. Analog FM exciter block diagram.

permitting wide deviation of the carrier frequency at baseband modulating frequencies. This implies that a PLL system behaves like an audio high-pass filter where higher modulating frequencies are ignored by the control loop while lower frequencies are considered errors in the average frequency and are tracked out by the loop. An added advantage of the PLL is the ability to synthesize the desired frequency from a single reference oscillator, thereby eliminating the need to change crystals when changing the frequency of the exciter.

The block diagram shown in Fig. 4 includes the key elements in the PLL. The output of the modulated oscillator operating at the carrier frequency is digitally divided down to a frequency of a few kilohertz or even less, called the comparison frequency. Likewise, the reference crystal oscillator is

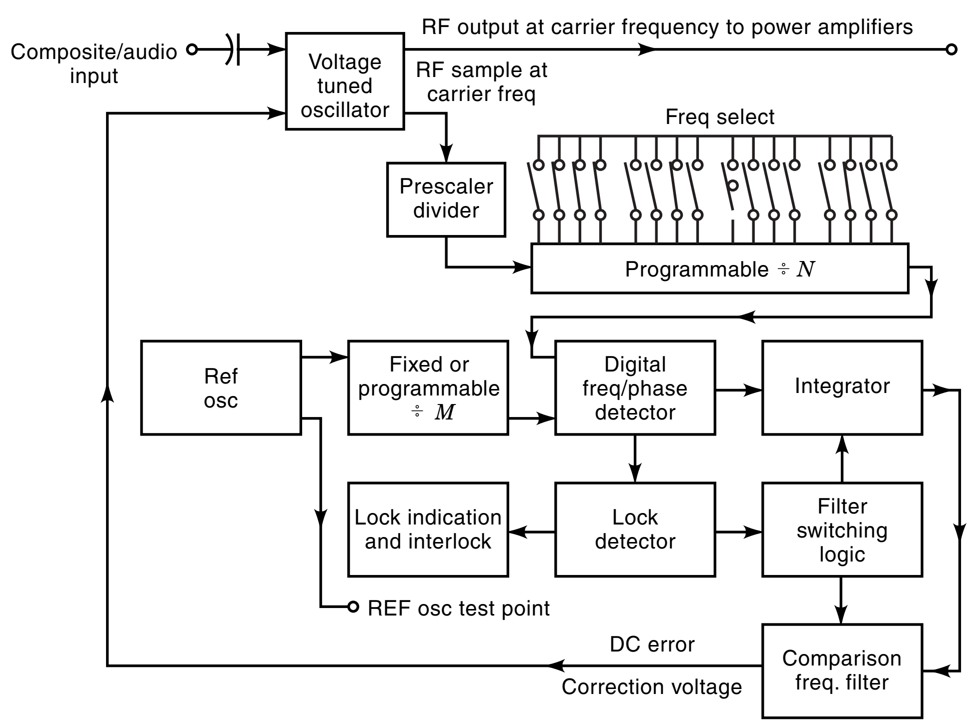


Figure 4. Phase-locked-loop frequency synthesizer.

also digitally divided down to the reference frequency. The two frequencies are compared in a digital phase/frequency detector to develop an error voltage which corrects the carrier frequency of the modulated oscillator. The reason for dividing the modulated oscillator frequency so many times is to reduce the modulation index enough to limit the peak phase deviation at the reference frequency to a value that will not exceed the linear range of the phase/frequency detector. If the linear range is exceeded, the loop will lose lock. This is why some exciters may lose AFC lock in the presence of low-frequency modulation components (2).

The phase detector output is integrated and low-pass filtered to remove the comparison frequency and all other frequency components above a few hertz so that the AFC circuit does not try to track out low-frequency modulation. Some FM exciters use a dual-speed PLL to keep the loop-turnover frequency low enough to maintain good amplitude and phase response at 30 Hz, while also providing quick lock-up time. The PLL error correction circuitry must respond quickly during the initial frequency scan of the FM band to achieve lock-up to the precision reference oscillator in a few seconds. The loop bandwidth is wide during acquisition and lock-up. After lock is achieved, the bandwidth is reduced to provide the optimum modulation characteristic.

The reference oscillator is usually temperature compensated and requires no warm-up to maintain ± 3 PPM or better accuracy over the operating temperature range. 10 MHz is often selected as the reference frequency for convenient comparison to international frequency standards.

Digital FM Exciter using Direct Digital Synthesis

A new technology called direct digital synthesis (DDS) eliminates the need for a phase-locked loop (PLL) in the FM modulation process by directly synthesizing the carrier frequency, including FM modulation, from a sine wave look-up table in a programmable read only memory device (PROM) operating in conjunction with a digital phase accumulator and a fast digital to analog converter (DAC). When this technique is combined with digital signal processing (DSP) technology, the entire process of generating stereo baseband with SCAs (Subsidiary Communications Authorization) and FM modulating this baseband information onto the RF carrier can be done entirely in the digital domain. The cost-to-performance ratio of DDS/DSP technology has made it competitive with the analog technology in present exciters (3). The full benefit of DDS/DSP technology requires digital transmission of audio information as an uncompressed, digital, bit stream all the way from the digital audio source through a digital console, digital audio processing, and an uncompressed, digital, studio-to-transmitter link (STL) to the AES3 digital input port of the DSP/DDS exciter (4). This same technology is used in the fully digital audio broadcast (DAB) services with the European Eureka 147 (EU-147) transmission standard and other technical standards presently implemented worldwide (5).

Direct Digital Synthesis of the FM Waveform

Direct digital synthesis (DDS) is a technique whereby the completely modulated FM waveform is generated totally in the digital domain. As digital modulation is an inherently linear process, no predistortion is required. The FM signal generated by a DDS device has extremely low noise and distortion, for true 16-bit digital audio quality (-96 dB FM signal-to-noise ratio and 0.0016% harmonic distortion for ± 75 kHz deviation and 75 μs preemphasis/deemphasis).

The current generation of DDS exciters use a 32-bit numerically controlled oscillator (NCO). The basic resting frequency of the NCO is set by a 32-bit tuning word. Frequency modulation occurs when modulation data varies the structure of the tuning word data within the phase accumulator section of the NCO. The modulated output of the NCO is converted to analog FM, upconverted, filtered, and amplified to become the RF excitation for a conventional FM broadcast transmitter RF amplifier chain. A block diagram of a DDS digital FM exciter is shown in Fig. 5.

DDS FM exciters also eliminate several basic limitations found in analog exciters using direct FM via the modulation of a voltage-controlled oscillator (VCO). Very low audio frequencies must be filtered from program signals to avoid affecting the automatic frequency control (AFC) circuits of the analog exciter, which see very low modulating frequencies as an off-frequency condition that needs correction. A DDS-based FM exciter has no such limitation, and modulation response extends virtually to dc (zero hertz). These lower octaves of program material are important for sonic realism and to preserve the phase correlation existing in the original program.

Digital Modulator

The digital modulator, which is the heart of a digital exciter, utilizes a 32-bit NCO to digitally generate the completely modulated FM waveform. Other supporting circuitry including a digital peak detector drives the front panel modulation display. The block diagram in Fig. 6 shows the functional subsystems of a digital modulator.

The input to the digital modulator is fed modulation data in the format and at the clock rate required. This data represents the stereo baseband created in the digital input module which contains the DSP digital stereo generator and subcarrier input circuitry.

The Digital modulator module includes a precise, digital peak detector to provide the drive for a peak modulation display. This circuit is driven by the same data as the NCO modulator. Therefore, the modulation indication on the front panel of the digital exciter has very high accuracy to within 0.25% of the true FM deviation value at any modulation index or frequency.

The output of the NCO is D/A converted to a precise, conventional FM modulated signal at an intermediate frequency and band-pass filtered to remove the images produced in the DDS process (6).

Digital Input and DSP Stereo Generator

DDS techniques are compatible with standard AES3 digital audio and provide the last link in maintaining a 100% digital audio path from the program source through the generation of the modulated FM carrier with no intervening A/D or D/A conversions to add noise and distortion. The DDS exciter includes a built-in DSP stereo generator to convert the incoming AES3 digital stereo into the digital stereo modulation data needed for the NCO to generate FM stereo. Most DDS exciters also offer as an option a high-quality A/D converter that will convert the analog baseband output of an analog

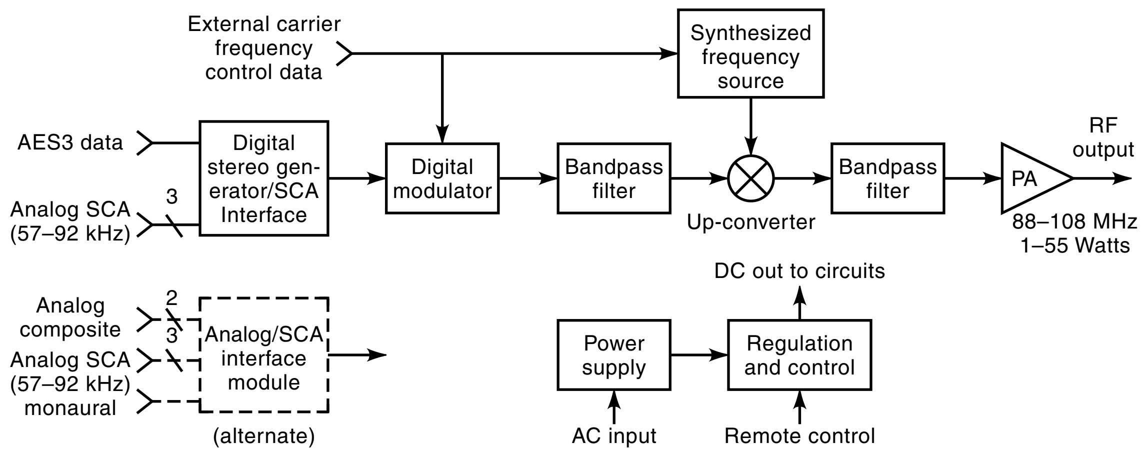


Figure 5. Harris DIGIT-CD DDS digital FM exciter.

stereo generator into the digital data format needed by the NCO.

Auxiliary signals, such as SCA (Subsidiary Communications Authorization) and RBDS/RDS (Radio Broadcast Data Service), are accepted as modulated analog waveforms from external devices, then A/D converted to digital data and applied to the NCO for simultaneous transmission with normal stereo program material. Figure 7 shows the major signal blocks in the digital input module (7).

The digital input module provides the interface, monitoring, and control circuitry for these functions:

1. Accepts standard AES3 stereo digital data at any rate from 20.8 kHz to 56 kHz, normally 32 kHz. Rate conversion is automatic (self-clocking).
2. Accepts analog subcarriers in the range of 57 kHz to 92 kHz.
3. Converts all input signals to the composite digital format needed by the digital modulator.
4. Provides a 19 kHz output to synchronize an external RBDS/RDS generator.
5. Supplies digital composite limiting using *look ahead* digital technology.
6. Provides gain adjustment for SCA, overall deviation, limiting level, and pilot level.
7. Provides a switchable LED display showing either total peak deviation or limiting level.
8. Allows user selection of program channel mode (AES3/EBU or analog) (Audio Engineering Society/European Broadcast Union Digital Audio Transport Standard), preemphasis on/off/time-constant, mono on/off, and data error detection mode.

Exciter Metering

Metering of important operating parameters can be provided by a combination of analog metering and a digital LED display. Steady-state parameters are usually selected by a multiposition switch and displayed on an analog or LCD multimeter. Typical steady-state functions include regulated, preregulated, and unregulated supply voltages; the AFC control voltage; RF power amplifier collector voltage and current; forward output power; and reflected power.

A color-coded peak reading display is usually provided to constantly monitor the peak FM deviation. A high-speed peak detector gives accurate peak readings on signals from dc to 100 kHz. A one-shot multivibrator circuit provides a clear in-

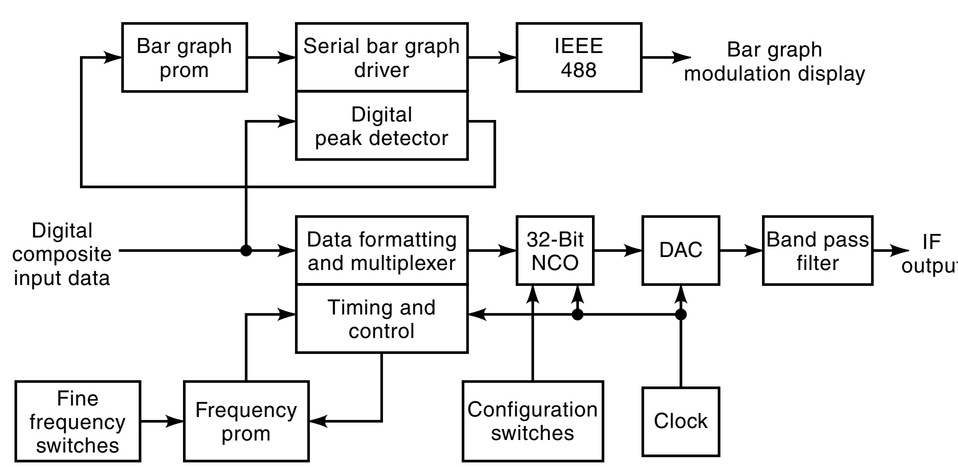


Figure 6. Digital modulator in the Harris DIGIT-CD FM exciter.

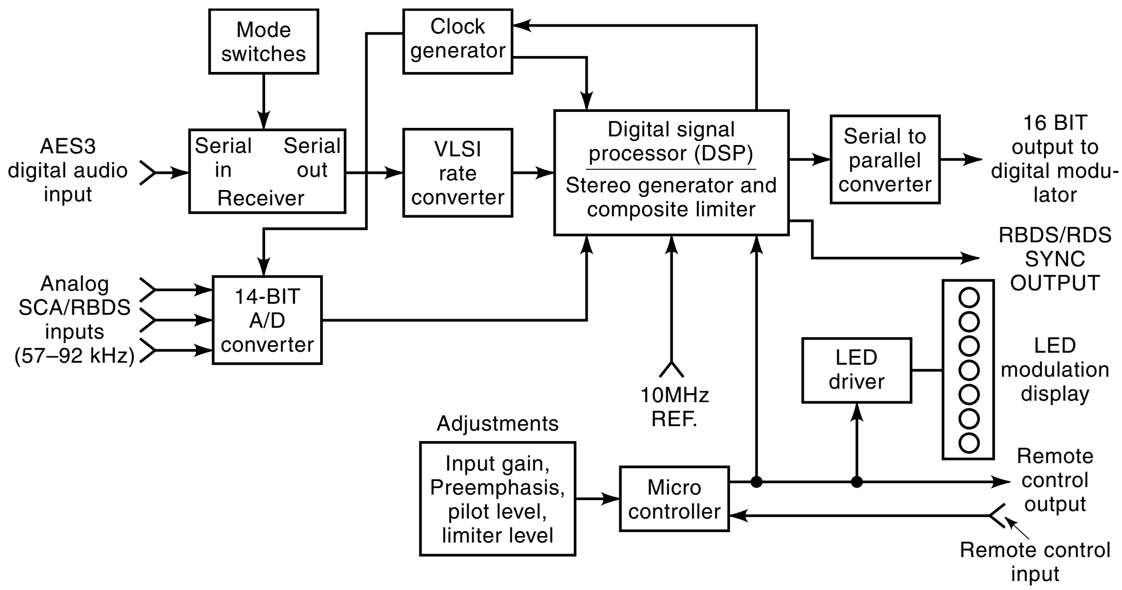


Figure 7. Harris DIGIT-CD digital input and DSP stereo generator.

dication of short transient peaks exceeding 100% modulation. Digital exciters directly read the peak values of the modulation data producing the FM deviation in the NCO.

Exciter Output Stage

The broadband RF amplifier in the exciter amplifies the output of the modulated oscillator from a power level of a few milliwatts up to an output level in the range of 5 W to 50 W. The output stage is usually protected against damage by an infinite VSWR at any phase angle.

The typical RF amplifier is designed to have a bandwidth of at least 20 MHz, using successive broadband impedance matching sections for each stage. Each group of matching sections consists of microstrip or lumped elements.

The broadband performance of the RF amplifier eliminates the need for adjustments to any particular frequency within the FM band. The exciter output is transparent to the signal generated by the modulated oscillator, and the amplifier stability is enhanced under varying load conditions.

A microstrip directional coupler is often incorporated in the RF amplifier output network. This coupler supplies information to the exciter control circuitry which provides automatic control of power output level and provides protection against operation under high VSWR conditions.

All current generation FM exciters will produce at least 50 W of RF output so they can be used as complete transmitters for educational stations with the addition of a harmonic filter to the output. For higher power levels, the exciter is used to drive an external power amplifier.

RF POWER AMPLIFIERS

The remainder of the FM transmitter consists of a chain of power amplifiers, each having from 8 dB to 20 dB of power gain. Ideally, the transmitter should have as wide a bandwidth as practical with a minimum of tuned stages. Broadband solid-state amplifiers are preferred to eliminate tuned networks in the RF path. Higher powered transmitters in the multikilowatt range may use multiple tube stages, each with fairly low gain, such as in the grounded-grid configuration or a single grid-driven PA stage with high gain and efficiency. The cost, redundancy, and wide bandwidth benefits of solid-state transmitters have made them attractive at power levels up to 20 kW. At higher power levels beyond 20 kW, the lower cost per watt of high-power, single-tube transmitters still make them attractive even though the modulation performance and reliability are less than those of a solid-state transmitter. Design improvements in tube-type power amplifiers have concentrated on improving bandwidth, reliability, and cost effectiveness whereas design improvements in solid state amplifiers have focused on continuous cost reduction to make them competitive with tube technology at ever increasing power levels (8).

RF Power Amplifier Performance Requirements

The basic function of the power amplifier is to amplify the exciter output to the authorized transmitter power output level. Most of the overall performance characteristics of the transmitter are determined by the exciter, but a few are established or affected by the power amplifier characteristics:

1. The RF output level at harmonics of the carrier frequency is almost completely a function of the attenuation provided by the output matching circuit and output low-pass/notch filters. The FCC limit in decibels is [43 dB + 10(log watts) dB] or 80 dB whichever is less (73 dB for 1 kW output or 80 dB for 5 kW and higher).
2. The major source of asynchronous AM noise usually originates in the last power amplifier stage. The FCC limit is 50 dB below 100% equivalent AM modulation.
3. The RF power output control system which must keep the output within +5% and −10% of authorized output is usually achieved in the final power amplifier.

4. Inadequate power amplifier RF bandwidth, particularly with respect to phase linearity (constant time delay) across the signal bandwidth, can reduce stereo separation and cause SCA crosstalk.
5. The presence of standing waves on the transmission line to the antenna may also interact with the power amplifier to cause degraded stereo separation and SCA crosstalk.

The power amplifier should provide trouble-free service and be easy to maintain and repair. Good overall efficiency is also desirable to reduce the primary power consumption and heat load released into the transmitter room.

Power Amplifier Bandwidth Considerations

As mentioned earlier, the FM signal theoretically occupies infinite bandwidth. In practice, however, truncation of the insignificant sidebands (typically less than 1% of the carrier) makes the system practical by accepting a certain degree of signal degradation. The transmitter power amplifier bandwidth affects the modulation performance. Available bandwidth determines the amplitude response and group delay response. There is a tradeoff involved between the bandwidth, gain, and efficiency in the design of a power amplifier (9,10).

The bandwidth of an amplifier is determined by the load resistance across the tuned circuit and the output or input capacitance of the amplifier. For a single-tuned circuit, the bandwidth is proportional to the ratio of capacitive reactance to resistance:

$$\mathrm{BW} \propto \frac{K}{2\pi f R_L C} = \frac{K X_C}{R_L} \tag{1}$$

where:

BW = bandwidth between half-power points (BW3)
K = proportionality constant
R_L = load resistance (appearing across tuned circuit)
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
X_C = capacitive reactance of C
f = carrier frequency (11).

The load resistance is directly related to the RF voltage swing on the tube element. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

EFFECTS OF CIRCUIT TOPOLOGY AND TUNING ON FM MODULATION PERFORMANCE

FM broadcast transmitter RF power amplifiers are typically adjusted for minimum synchronous AM (incidental amplitude modulation with FM modulation) which results in symmetrical amplitude response. This will assure that the transmitter's amplitude passband is properly centered on the FM channel. The upper and lower sidebands will be attenuated equally or symmetrically which is *assumed* to result in optimum FM modulation performance. This will be true if the RF power amplifier circuit topology results in simultaneous symmetry of amplitude and group delay responses.

Actually, symmetry of the group delay response has a much greater effect on FM modulation distortion than the amplitude response. Tuning for symmetrical group delay will cause the phase/time delay errors to affect the upper and lower sidebands equally or symmetrically. The group delay response is constant if the phase shift versus frequency is linear. All components of the signal are delayed equally in time, but no phase distortion occurs.

The tuning points for symmetrical amplitude response and symmetrical group delay response usually do *not* coincide, depending on the circuit topology. Therefore, simply tuning for minimum synchronous AM (symmetrical amplitude response) does not necessarily result in the best FM modulation performance (12).

Measurements taken on typical FM transmitters as well as computer simulations show that tuning the RF power amplifier for symmetrical group delay response results in minimum distortion and cross talk. Group delay response asymmetry causes higher FM modulation distortion and cross talk than amplitude response asymmetry. The transmitter should be tuned for symmetrical group delay response which results in the best FM modulation performance rather than symmetrical amplitude response which results in minimum synchronous AM (13).

Power Amplifier Output Source Impedance

At the milliwatt levels used in RF test equipment, it is customary to provide 50 Ω source and load impedances at both ends of a coaxial transmission line. This approach minimizes any reflections on the line because both the transmitter (source) and the termination (load) absorb reflected energy. A 50 Ω source impedance is usually provided by placing a 50 Ω build-out resistor in series with a low-impedance voltage source (Thevenin equivalent). The closed circuit voltage with this configuration is exactly one-half of the open circuit voltage, meaning that half of the total available RF power is dissipated in the source resistance. The best possible efficiency for this system is 50% assuming that the voltage source is 100% efficient without the source resistance.

It becomes obvious that, although an FM transmitter is designed to drive a 50 Ω load, it does not itself have an output source impedance of 50 Ω. To achieve high efficiency, the transmitter must have a very low output source impedance so that nearly all of the power is delivered to the load. The plate dissipation indirectly represents some of the power lost within the low source resistance. Because the low source impedance of the transmitter provides a mismatch to reflected power from the load, this power is almost totally reflected back from the transmitter output stage toward the load again.

Intermediate Power Amplifiers

The intermediate power amplifier (IPA) is located between the exciter and the final amplifier in higher power transmitters that require more than about 50 W of drive to the final amplifier. The IPA may consist of one or more tubes or solid-state amplifier modules.

Most of the newer design, high-power transmitters with a tube in the final amplifier require between 150 W and 600 W of drive. This permits the use of solid-state, wideband, power amplifier modules to boost the exciter's power up to the level required to drive the grid of the final tube.

Interstage Coupling Circuits

The separate IPA output circuit and the final amplifier input circuit are often coupled together by a coaxial transmission line. Impedance matching is usually accomplished at either end by one of the configurations shown in Fig. 8.

The interconnecting transmission line between the coupling circuits should be properly matched to avoid a high VSWR. Directional wattmeters are normally placed in the line to measure forward and reflected power from which a standing wave ratio can be established. The VSWR is established by the match at the load end of the transmission line.

Solid-state RF power devices possess a very low load impedance at the device output terminal, so that an impedance transformation that goes through the 50 Ω intermediate impedance level is required to couple these devices into the relatively high impedance of the final amplifier grid circuit. Therefore, virtually all solid-state IPA systems have a 50 Ω impedance point within the system that can be used to feed the antenna in an emergency.

Solid-State RF Power Amplifier Systems

A solid-state RF power amplifier almost always consists of a system of individual amplifier modules combined to provide the desired power output. Following are the advantages of using several lower power modules instead of a single high-power amplifier:

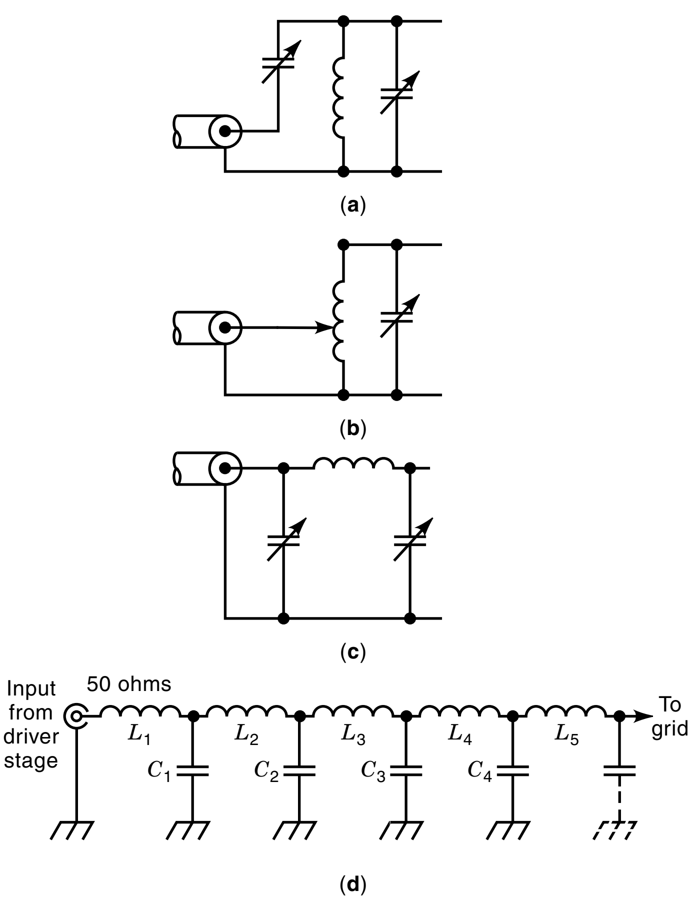


Figure 8. Interstage RF coupling circuits.

1. Redundancy is provided by isolating the input and output of each module, permitting uninterrupted operation at reduced power if one or more of the modules fails.
2. The ability to repair or replace failed modules without having to go "off the air."
3. More effective cooling of each power device junction by splitting the concentration of heat to be dissipated into several areas instead of one small area.
4. Better isolation between the amplifier modules and the input circuit of the final power amplifier or antenna is provided by the combiner/isolator.
5. Redundant power supplies and air cooling systems for each module improve overall reliability.

Each RF power amplifier module consists of one or more solid-state devices with broadband impedance transformation networks for input and output matching. A new generation of class "*C*" MOSFET devices permits the design of broadband amplifier stages with both high efficiency and the wide bandwidth necessary to cover the FM broadcast band.

The input impedance to the solid-state device is always lower than the desired 50 Ω input impedance, so a broadband impedance transformation scheme is required. This is usually accomplished by a combination of coaxial baluns and push-pull coaxial line sections cross-coupled to provide 4 : 1 or higher transformation ratios over the FM band.

By operating two devices in push-pull, the input impedance (differential) is double that of a single ended circuit, and the suppression of even order harmonics is enhanced. Two devices fed in this manner also provide some degree of redundancy within the module itself because partial RF output may be obtained when one device fails. Similarly, the low output impedance of these solid-state devices can be transformed up to the desired 50 Ω module output impedance where combining occurs. Figure 9(a) illustrates a simplified schematic of a broadband, 350 W, MOSFET, RF amplifier module utilizing the push-pull configuration. Figure 9(b) is a photograph of this RF amplifier module (14).

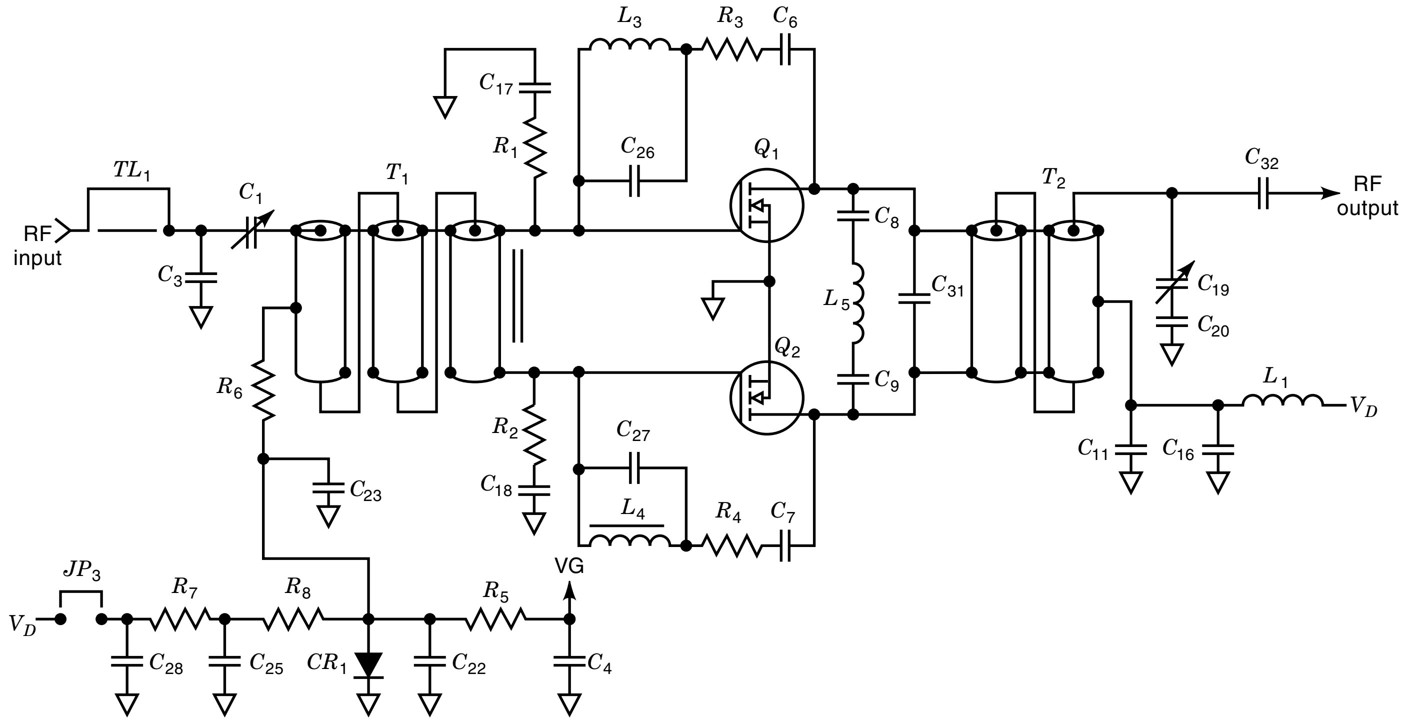
Solid-State Amplifier Splitting and Combining

The following are two frequently used types of splitting/combining schemes are:

1. A 90° hybrid splitter or combiner ($N - 1$ hybrids are required to split or combine N inputs) (see section on transmitter output combining).
2. A Wilkinson *N-way* in-phase splitter or combiner.

Either type of splitter/combiner must provide isolation between the individual power amplifier modules and low-loss splitting or combining of the total power.

The cascaded 90° hybrid system shown in Fig. 10 provides double isolation between the power amplifiers and the load by first combining the two pairs of amplifiers and then combining the outputs of the first two combiners. A portion of the reflected power, caused by a mismatch at the output, will be dissipated in the reject loads so that the power amplifier modules will operate into a lower VSWR than exists at the output. The unbalanced 50 Ω reject loads are accessible for monitoring of reject load power which is useful in determining the balance of the system.



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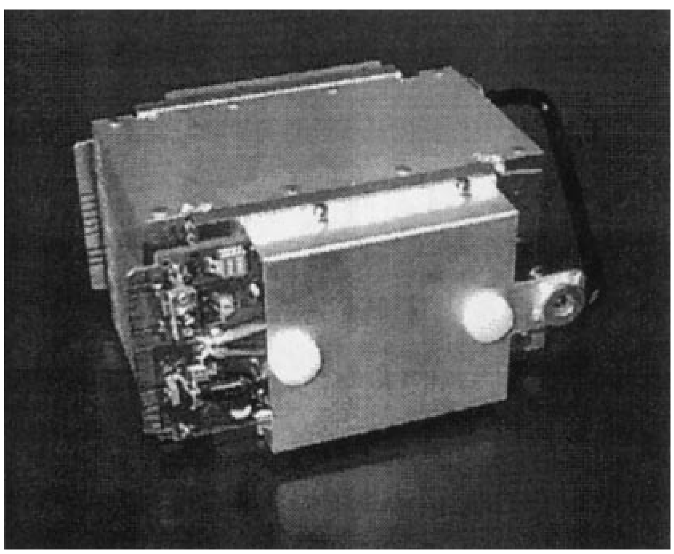

(**b**)

Figure 9. (a) Schematic of broadband, 350 W, MOSFET, RF power amplifier module. (b) Broadband, 350 W, MOSFET, RF power amplifier module.

The Wilkinson system shown in Fig. 11 is a simple and effective way to split and combine modules operating in phase but usually requires a balanced reject load which makes reject power measurements more difficult. By adding additional coaxial balun sections to the Wilkinson, it is possible to use unbalanced reject loads (15). This configuration is called Wilkinson-Gysel.

ADAPTIVE CONTROL OF THE COMBINER CONFIGURATION

Both the 90° hybrid and the Wilkinson combining systems require resistive RF power reject loads to provide isolation between the amplifier modules in the event that one or more of the modules fail. A portion of the RF power from the remaining modules is wasted in the reject loads instead of being

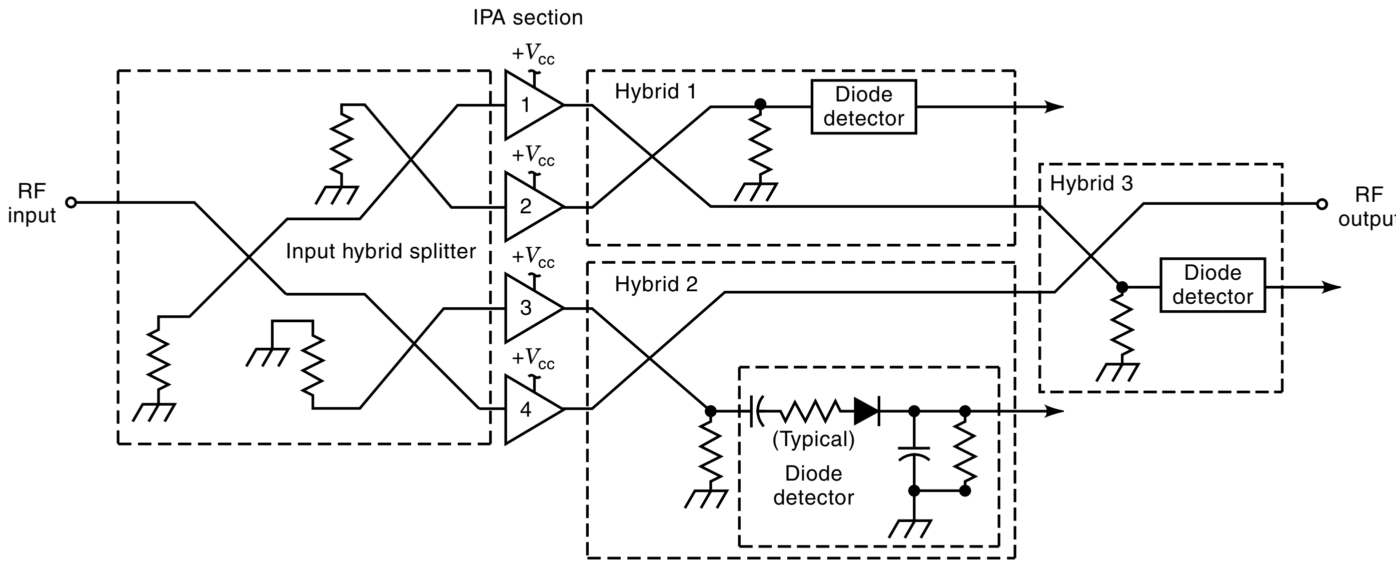

Figure 10. Cascaded 90° hybrid combiner.

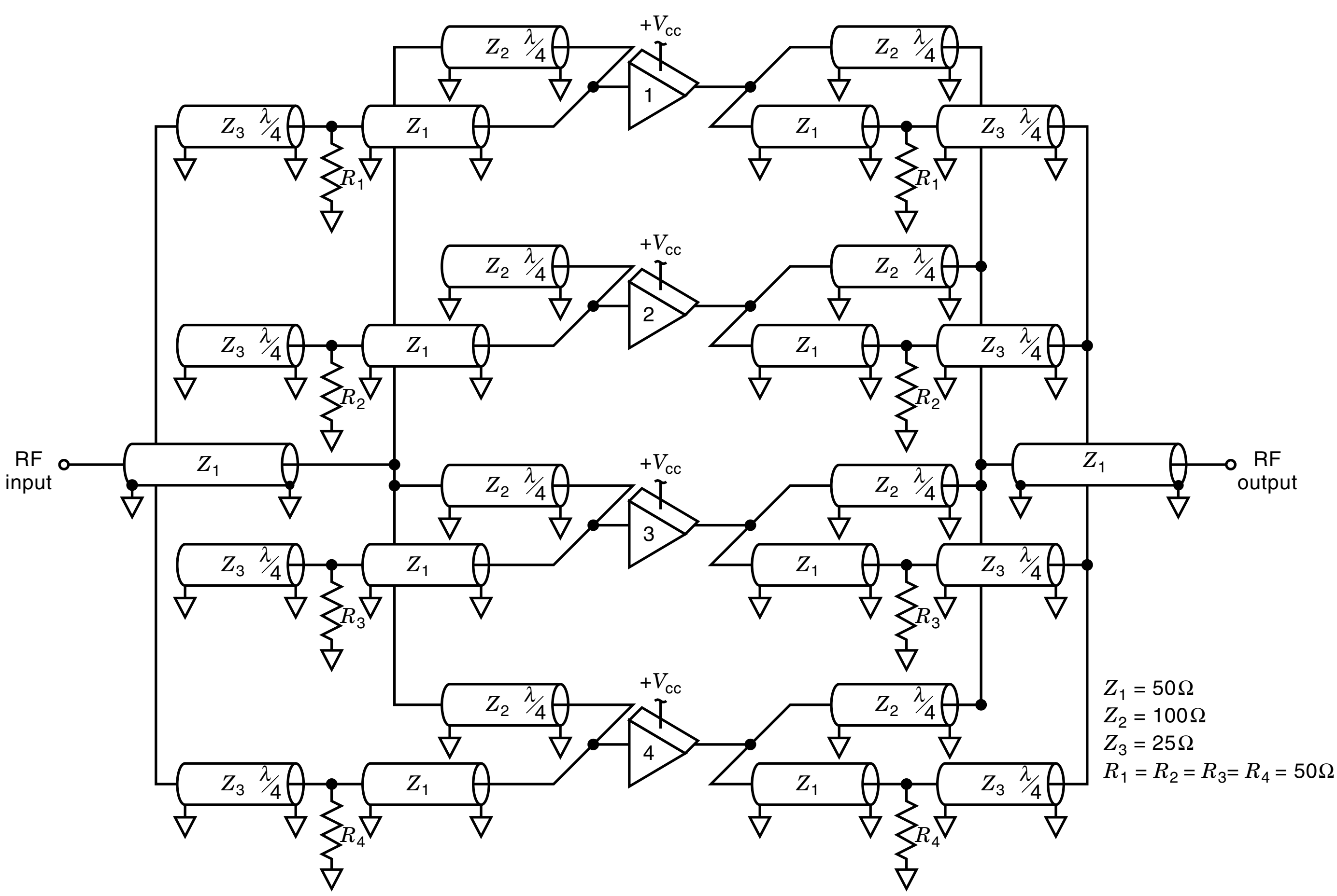

Figure 11. Wilkinson–Gysel in-phase splitting/combining system with unbalanced reject loads.

delivered to the output. Recent developments have made it possible for a microcomputer to monitor the degree of imbalance in the system and adaptively change the configuration of the combiner to losslessly compensate for the failure of one or more power amplifier modules. This is accomplished by having the microcomputer substitute the appropriate reactances in place of the resistive reject loads to maintain enough isolation for the remaining power amplifiers to work efficiently. This technique is used in the Harris "*Z*" plane combiner.

Because most splitter/combiner systems are designed around a 50 Ω input and output impedance level, these systems can be easily used as a low-power standby transmitters by routing the output to the antenna system. An RF low-pass filter (LPF) is required only when directly feeding the antenna system. The harmonic suppression of the IPA is not as critical when driving a nonlinear power amplifier that also generates harmonics, because this stage will have its own LPF.

SOLID-STATE FM BROADCAST TRANSMITTERS

The techniques used to construct IPA systems can also be used to construct a completely solid-state transmitter using arrays of combined modules for the final output stage. An additional RF low-pass filter is usually required to meet FCC emission requirements.

Advantages of Solid-State Transmitters

The primary advantages of a solid-state transmitter are the built-in amplifier and power supply redundancy, superior FM modulation performance, the ability to cover the entire FM band without the need for retuning, and elimination of tube replacement costs. A tubeless transmitter is nearly maintenance free.

Solid-State Transmitter Design Considerations

Several manufacturers offer solid-state FM broadcast transmitters with power outputs ranging from 100 W up to 20 kW, but present economic factors still favor the single-tube FM transmitter for power levels above 20 kW. For a solid-state transmitter to be competitive in cost and power consumption with a single tube transmitter, the efficiency of the solid-state RF power amplifiers and combining system have to approach the 80% efficiency obtainable from tube type RF amplifiers. This high efficiency has recently been achieved with MOSFET solid-state devices at VHF frequencies.

Recent solid-state designs have provided higher efficiencies, up to 80% dc to RF efficiency at the MOSFET device level and over 62% overall efficiency, from ac line in to RF output. This is actually better ac to RF efficiency at the 5 kW level than a typical single-tube transmitter.

Some solid-state designs have added a few percent to their overall ac to RF efficiency by optimizing their RF circuits over narrowband sections of the FM band. This approach is beneficial to the user who is certain that there will be no need to change the transmitter's frequency or who is prepared to provide the transmitter modifications needed to do so.

Trends in the newest solid-state FM transmitters are to supply redundant RF, power supply, and control circuits so as to keep the transmitter on the air at reduced power in the event one or more components should fail. Identical and interchangeable IPA and PA modules offer additional redundancy. RF modules that can be removed and inserted in an operating transmitter also provide the advantage of not requiring an off-air period for some maintenance services.

Solid-state transmitter layouts with direct, cable-free connection of the RF modules to the RF combiner have also been introduced and further enhance transmitter reliability and stability. Another enhancement provided in some current solid-state FM transmitters is an advanced, microprocessor-based, control system that monitors detailed parameters within the transmitter and provides *intelligent control* of the transmitter system, including the RF combiner, so as to maximize output power and minimize reject load power under various combinations of active and inactive modules.

Figure 12 shows a block diagram of a 5 kW solid-state transmitter.

Vacuum-Tube Power Amplifier Circuits

The amplitude of an FM signal remains constant with modulation so that efficient, nonlinear, Class *C,* amplifiers can be used.

FM broadcast vacuum-tube power amplifier circuits have evolved into two basic types. One type uses a tetrode or pentode tube in a grid-driven circuit whereas the other uses a high-μ triode in a cathode-driven (grounded grid) circuit.

Cathode-Driven Triode Amplifiers. The high-μ triodes being used in cathode-driven (grounded-grid) FM amplifiers were originally developed for linear SSB amplifiers. Their characteristics are well adapted to FM broadcast use because the circuit is very simple and no screen or grid bias power supplies are required. Figure 13 shows the basic circuit configuration. In this case, the grid is connected directly to chassis ground. Dc grid current is the difference between dc cathode current and dc plate current. The output tank circuit is a shorted coaxial cavity which is capacitively loaded by the tube output and stray circuit capacitance. A small capacitor is used for trimming the tuning and another small variable capacitor is used to adjust the loading. A π-network matches the 50 Ω input to the tube cathode impedance.

The triodes are usually operated in the less efficient, class "*B*" mode to achieve maximum power gain, which is on the order of 20 (13 dB). They can be driven into high-efficiency, class "*C*" operation by providing negative grid bias. This increases the plate efficiency, but also requires increased drive power.

Most of the drive power into a grounded-grid amplifier is fed through the tube and appears in the stage's output. This increases the apparent efficiency so that the efficiency factor given by the manufacturer may be higher than the actual plate efficiency of the tube. The true plate efficiency is determined by dividing the output power by the total input power, which includes both the dc plate input power ($I_p \times E_p$) and the RF drive power. Because most of the drive power is fed through the tube, any changes in loading of the output circuit also affect the input tuning and driver stage.

There is RF drive voltage on the cathode (filament) of the tube, so some means of decoupling must be used to block it from the filament transformer. One method employs high-cur-

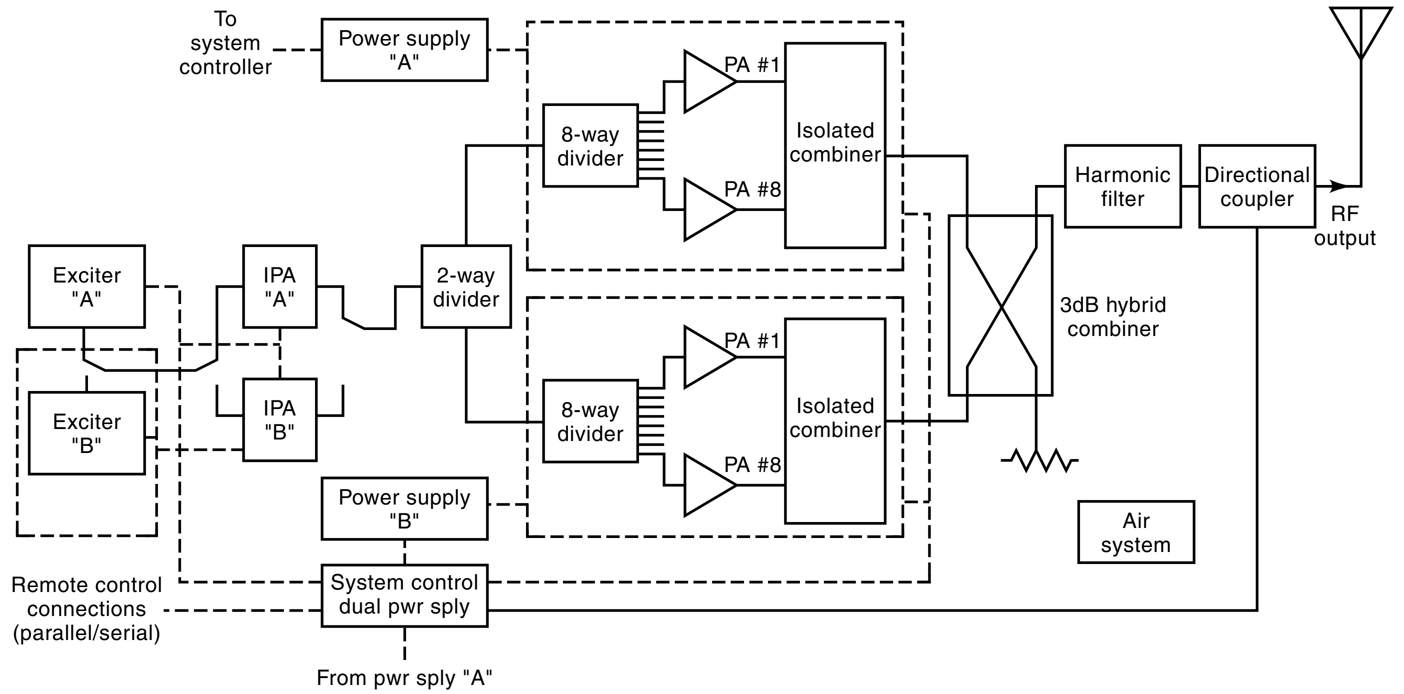

Figure 12. Harris *Z* 5 kW solid-state FM transmitter.

rent RF chokes because the inductance can be very low at this frequency range. The other commonly used method feeds the filament power through the input tank circuit inductor.

Cathode-driven stages are normally used only for the higher power stages. The first stage in a multitube transmitter is nearly always a tetrode because of its higher power gain.

Grid-Driven Tetrode and Pentode Amplifiers. Transmitters with tetrode amplifiers throughout usually have one less stage than those with triodes. Because tetrodes have higher power gain, they are driven into class "*C*" operation for high plate efficiency. Against these advantages is the requirement for neutralization, along with screen and bias power supplies.

Figure 14 shows a schematic of a grid-driven tetrode amplifier. In this example, the screen is operated at dc ground potential and the cathode (filament) is operated below ground by the amount of screen voltage required. This is called grounded-screen operation. It has the advantage that stability problems due to undesired resonances in the screen-bypass capacitors are eliminated. With directly heated tubes, it is necessary to use filament-bypass capacitors. During grounded-screen operation, these bypass capacitors need a higher breakdown voltage rating because they have the dc screen voltage across them. The filament transformer must have additional insulation to withstand the dc screen voltage.

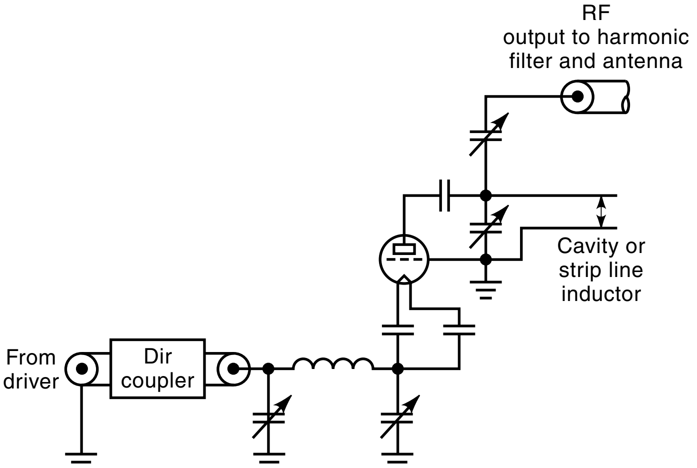

Figure 13. Cathode-driven, triode, power amplifier.

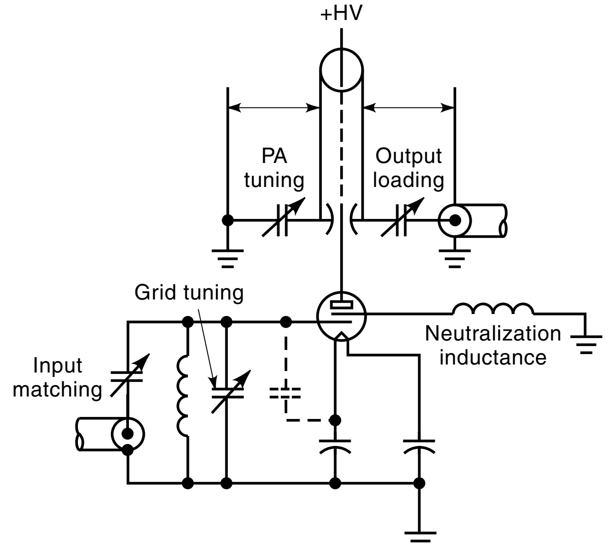

Figure 14. Grid-driven, grounded-screen, tetrode, power amplifier.

The screen power supply provides a negative voltage in series with the cathode to ground and must have the additional capacity to handle the sum of the plate and screen currents. A coaxial cavity is used in the output circuit so that the circulating current is spread over large surfaces to keep the losses very low. This cavity is a shorted quarter-wavelength transmission line section which resonates the tube's output capacitance. The quarter-wavelength cavity is actually shorter than a physical quarter-wavelength due to the electrical loading effect of the tube's output capacitance across the open end of the transmission line. The length is preset to the desired carrier frequency, and then a small value variable capacitor is used to trim the system to resonance. Capacitive output coupling is used to match from the high RF voltage point to the 50 Ω transmission line.

The 50 Ω input is capacitively coupled into the grid circuit inductor to provide the correct impedance match.

Pentode amplifiers have even higher gain than their tetrode counterparts. The circuit configuration and bias supply requirements for the pentode are similar to the tetrode because the third (suppressor) grid is tied directly to ground. The additional isolating effect of the (suppressor) grid eliminates the need for neutralization in the pentode amplifier (16).

Impedance Matching Into the Grid. The grid circuit is usually loaded (swamped) with added resistance. The purpose of this resistance is to broaden the bandwidth of the circuit by lowering the circuit Q and to provide a more constant load to the driver. It also makes neutralizing less critical so that the amplifier is less likely to become unstable.

Cathode or filament lead inductance from inside the tube through the socket and filament capacitors to ground can heavily load the input circuit. This is caused by RF current flowing from grid to filament through the tube capacitance and then through the filament lead inductance to ground. An RF voltage is developed on the filament which in effect causes the tube to be partly cathode-driven. This undesirable extra drive power requirement can be minimized by series resonating the cathode return path with the filament bypass capacitors or by minimizing the cathode-to-ground inductance with a specially designed tube socket containing thin-film dielectric *sandwich* capacitors for coupling and bypassing.

High-power, grid-driven, class "C," amplifiers require a swing of several hundred RF volts on the grid. To develop this high-voltage swing, the input impedance of the grid must be increased by the grid input matching circuit. Because the capacitance between the grid and the other tube elements may be 100 pF or more, the capacitive reactance at 100 MHz will be very low unless the input capacitance is resonated in parallel with an inductor. Figures 15(a) and 15(b) show two popular methods of resonating and matching into the grid of a high-power tube. Both methods can be analyzed by recognizing that the desired impedance transformation is produced by an equivalent L network.

In Fig. 15(a), a variable inductor L_{in} is used to raise the input reactance of the tube by bringing the tube input capacitance C_{in} almost to parallel resonance. Parallel resonance is not reached because a small amount of parallel capacitance C_p is required by the equivalent L network to transform the high impedance Z_{in} of the tube down to a lower value through the series matching inductor L_s. This configuration provides an L-network low-pass filter by using part of the tube's input capacitance to form C_p.

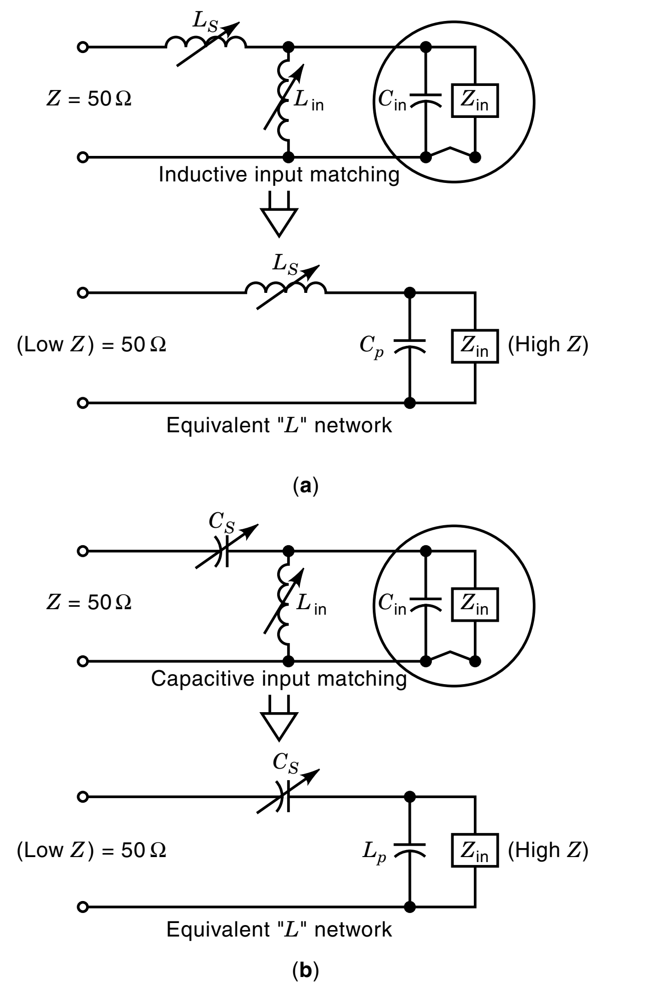

Figure 15. (a) Inductive input matching. (b) Capacitive input matching.

Figure 15(b) uses variable inductor L_{in} to take the input capacitance C_{in} past parallel resonance so that the tube's input impedance becomes slightly inductive. The variable series matching capacitor C_s forms the rest of the equivalent L network. This configuration is a high-pass filter.

Neutralization. Cathode-driven, grounded-grid amplifiers utilizing triodes do not require neutralization. It is necessary that the grid-to-ground inductance, both internal and external to the tube, be kept very low to maintain this advantage. Omission of neutralization allows a small amount of interaction between the output circuit and the input circuit through the plate-to-filament capacitance. This effect is not very noticeable because of the large coupling between the input and output circuits through the electron beam of the tube. Cathode-driven tetrodes have higher gain and therefore require some form of neutralization.

Grid-driven, high-gain tetrodes need accurate neutralization for best stability and performance. Self-neutralization can be accomplished very simply by placing a small amount of inductance between the tube screen grid and ground, usually in the form of several short, adjustable-length straps. The RF current flowing from plate to screen in the tube also flows through the screen lead inductance. This develops a small RF voltage on the screen, of the opposite phase, which cancels the voltage fed back through the plate-to-grid capacitance. This method of lowering the self-neutralizing frequency of the tube works only if the self-neutralizing frequency of the tube/socket combination is above the desired operating frequency before the inductance is added. Feedback neutralization utilizes a small coupling capacitor, usually in the form of a small plate located near the anode of the tube. The sample of the RF voltage from the anode intercepted by this plate is coupled through a 180° phase-shift network into the grid circuit. This technique has the advantage of providing neutralization over a very broad range of frequencies if implemented correctly, and stray reactances are minimized. Special attention must also be given to minimizing the inductances in the tube socket by integrating distributed bypass capacitors into the socket and cavity deck assembly. Pentodes normally do not require neutralization because the suppressor grid effectively isolates the plate from the grid.

Power Amplifier Output Circuits. Usually, the output circuit consists of a *high-Q* (low-loss) transmission line cavity, strip line, or a lumped inductor that resonates the tube output capacitance. A means of trimming the tuning and a means of adjusting the coupling to the output transmission line must also be provided by the output circuit. The tank circuit loaded Q is kept as low as practical to minimize circuit loss and to maintain as wide an RF bandwidth as possible.

The Power Amplifier Cavity. The vacuum-tube power amplifier is constructed in an enclosure containing distributed tank circuit elements for minimum loss. The efficiency of the PA depends on the RF plate voltage swing, the plate current conduction angle, and the cavity efficiency. The cavity efficiency is related to the ratio of the loaded to unloaded Q as follows:

$$N = 1 - \left(\frac{Q_L}{Q_U}\right) \times 100 \qquad (2)$$

where N is the efficiency in percent, Q_L is the loaded Q of cavity, and Q_u is the unloaded Q of cavity.

The loaded Q depends on the plate load impedance and output circuit capacitance. Unloaded Q depends on the cavity volume and the RF resistivity of the conductors due to skin effects. A high unloaded Q is desirable, as is a low loaded Q, for best efficiency. As the loaded Q goes up, the bandwidth decreases. For a given tube output capacitance and power level, the loaded Q decreases with decreasing plate voltage and increasing plate current. The increase in bandwidth at reduced plate voltage occurs because the smaller load resistance is directly related to the RF voltage swing (for the same power) on the tube element. For the same power and efficiency, the bandwidth can also be increased if the output capacitance is reduced. Power tube selection and minimization of stray capacitance are areas of prime concern when designing for maximum bandwidth.

The Quarter-Wavelength Cavity. The "quarter-wavelength" coaxial cavity is the compact and popular output circuit illustrated in Fig. 16. The tube anode is coupled through a dc blocking capacitor to a shortened "quarter-wavelength" transmission line. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a quarter-wavelength long. Plate tuning is accomplished either by adding end-loading capacitance at the high-impedance end of the line with a variable capacitor or by changing the position of the ground plane at the low-impedance end of the line. The plate-tuning capacitor may be a sliding or rotating plate near the anode of the tube. The center conductor of the transmission line (air exhaust chimney) is at dc ground whereas the anode of the tube operates at a high RF and dc potential. dc voltage is fed through an isolated "quarter-wavelength" decoupling network inside the chimney to the anode of the tube. The plate blocking capacitor prevents dc current flow from the anode into the chimney.

The Folded, Half-Wavelength Cavity. Another approach to VHF power amplification uses the reentrant, folded, "half-wavelength" cavity design illustrated in Fig. 17. The dc anode voltage is applied to the lower portion of the plate line through a choke at the RF voltage null point. The "half-wavelength" line is tuned by mechanically expanding or contracting the physical length of a flexible extension (bellows) on the end of the secondary transmission line stub, which is located concentrically within the primary transmission line (air exhaust chimney). Coarse frequency adjustment is accomplished by presetting the depth of the top secondary section of plate line into the tank cavity.

Other power amplifier configurations may use lumped components or hybrid combinations with distributed transmission line elements to achieve similar results. The discrete

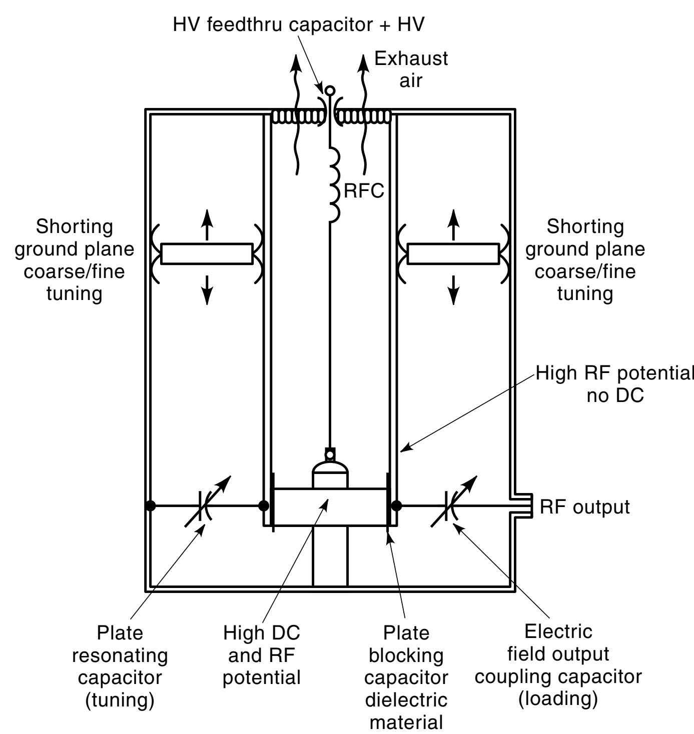

Figure 16. The quarter-wavelength cavity.

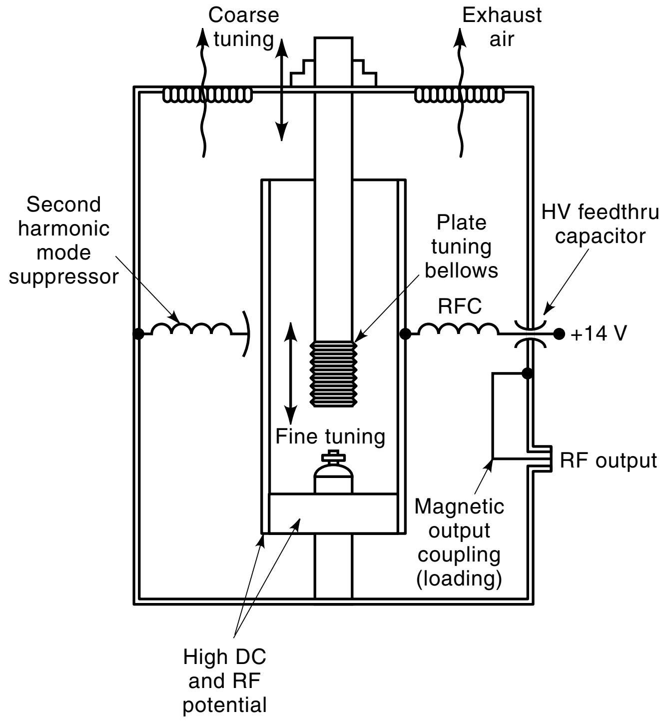

Figure 17. The folded, half-wavelength cavity.

circuit elements are chosen for their individual inductance or capacitance, instead of being operated in a purely "quarter-wavelength" or "half-wavelength" mode. Stray inductance and capacitance add to the component values resulting in the hybrid nature of these circuits.

The RF voltage and current distributions for the "quarter-wavelength" and the folded, "half-wavelength" cavities are shown in Fig. 18.

Regardless of the specific configuration, the output circuit must transform the high resonant plate impedance down to the output transmission line impedance of 50 Ω. The bandwidth of a transmission line cavity is optimized by choosing the highest characteristic impedance mechanically and electrically allowable.

Output Coupling. Power may be coupled from a "quarter-wavelength" cavity to the transmission line by a capacitive probe located near the high RF voltage point located at the anode end of the "quarter-wave" line as shown in Fig. 16. The amount of output coupling capacitance is determined by the RF power output required. The loaded Q of this circuit varies with the degree of capacitive coupling. Another method of coupling power from the "quarter-wavelength" cavity uses a tuned loop located near the grounded (high current) end of the line. In this case, the tuned loop operates both as an inductive and a capacitive pickup device. Power may be coupled from the "half-wavelength" line by an inductive loop located in the strong fundamental magnetic field near the center of the cavity, as shown in Fig. 17.

RF Output Low-Pass Filters. The high-efficiency, nonlinear RF power amplifiers used in FM broadcast transmitters generate significant amounts of energy on frequencies that are integral multiples (harmonics) of the desired fundamental frequency. The output circuit alone does not provide enough harmonic attenuation to meet FCC regulations. To comply with Part 73 of the FCC rules and regulations and to prevent interference to other services, a low-pass filter must be installed in the transmission line at the output of the transmitter. The FM band is narrow enough that one low-pass filter design can be used for any FM channel carrier frequency. These filters usually consist of multiple LC sections arranged so that frequencies within the FM band are passed with little attenuation (typically 0.1 dB or less) whereas frequencies above the FM band are highly attenuated (60 dB or more).

The most common type of filter in this application is called a *reflective* filter, meaning that the frequency components outside the passband are reflected back out of the filter toward the source because it provides a mismatch at these undesired

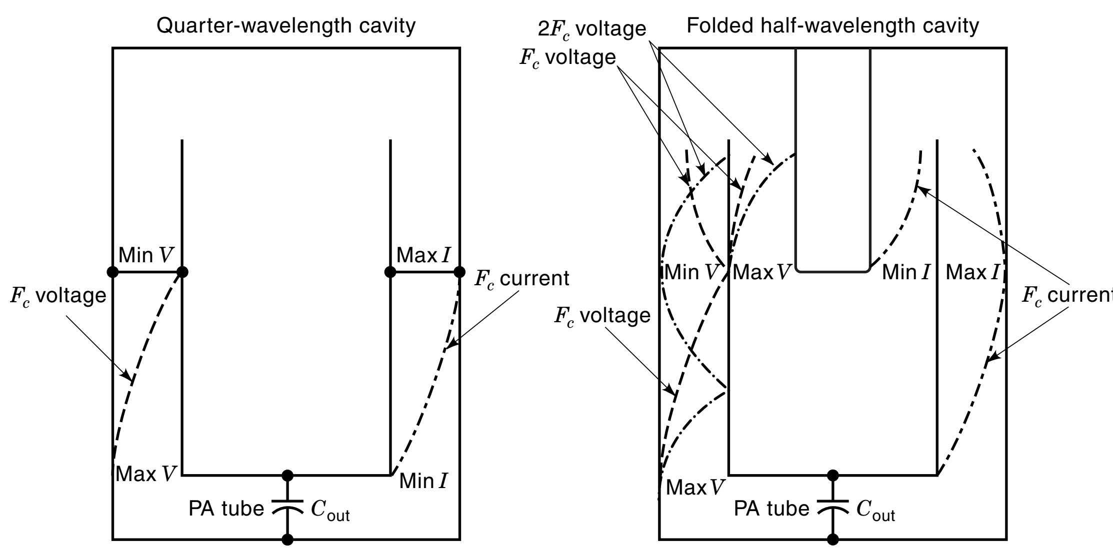

Figure 18. Cavity RF voltage and current distributions.

frequencies. The filter can be constructed using either *lumped* inductors and capacitors or by using a section of nonconstant impedance transmission line to form *distributed* inductors and capacitors. The filters designed for low-power transmitters often employ *lumped* elements (coils and capacitors) because they are compact and can be integrated into the transmitter cabinet. The distributed type of filter is most often used with high-power FM broadcast transmitters because of its simplicity, extreme ruggedness, and ability to handle higher power levels. The distributed filter does have the disadvantage of having larger physical dimensions than a similar lumped filter, which may necessitate mounting the filter outside of the transmitter cabinet. Figure 19 shows a cutaway view of a typical distributed low-pass filter. Note that the areas where the center conductor of the transmission line is smaller than that required for the input Z_0 are inductive, whereas the areas where the center conductor is larger in diameter are capacitive.

When two filters (such as the output cavity and the harmonic filter) are connected together by a transmission line, the total harmonic attenuation varies with interconnecting line length. The attenuation characteristics of the harmonic filter are specified for the condition where both the source and load impedances are equal to the desired transmission line impedance.

In actual use, the source impedance at the output of the tank circuit is much less than the 50 Ω load impedance presented by a properly terminated filter. At the operating frequency, the output impedance of the power amplifier and the input impedance of the low-pass filter become predominantly reactive at harmonic frequencies causing interaction between the two. If an unfortunate length of line is selected, the harmonic attenuation may be insufficient, and the transmitter tuning may be affected. This undesirable condition can be corrected by changing the line length by approximately one-quarter wavelength. At the operating frequency, the line length between the tank circuit and harmonic filter is usually supplied precut to a value known to be satisfactory by the transmitter manufacturer.

Harmonic Notch Filters. In some cases, a second harmonic notch filter is required in addition to the low-pass filter because the second harmonic component from the amplifier is high and the cutoff slope of the low-pass filter is not steep enough to provide sufficient second harmonic attenuation. The additional attenuation required (typically 30 dB) can be provided by a notch filter which places a short circuit across the transmission line at the second harmonic while providing a high impedance at the fundamental. A one-quarter wavelength (at the fundamental frequency), shorted coaxial stub is often used for this function. The second harmonic energy is primarily reflected back toward the power amplifier and to a lesser extent dissipated in the equivalent series resistance of the series tuned circuit formed by the stub. This shorted stub provides a very low impedance and a dc path from the center conductor of the transmission line to ground providing a separate, protective, advantage by shunting static discharges, such as lightning, to ground.

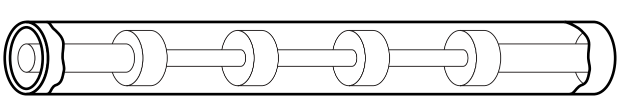

Figure 19. Cutaway view of a distributed low-pass filter.

COMBINED TRANSMITTERS

It is possible to combine the output of two RF power amplifiers for higher power levels. The important advantage is that the broadcast transmission is not interrupted if one amplifier fails. The radiated signal strength merely drops 6 dB until the failed amplifier is repaired and put back on the air. A dual amplifier system costs more than a single amplifier for a given total power output, but there are the economic advantages of reducing lost air time and eliminating the need for a separate standby transmitter. Automatic or manual output switching can be used to route the full power of the remaining amplifier directly to the antenna, reducing the loss in radiated power from 6 dB to 3 dB.

Two methods may be used to bypass the output-combining hybrid to allow 100% of the power of the remaining transmitter to be sent to the antenna if one transmitter of a combined pair should fail.

The first method uses three motorized switches (or patch panels) to bypass the 3 dB hybrid while connecting the operating transmitter directly to the antenna and the failed transmitter directly to the test load. This allows recovery of the 50% power lost in the reject load when one transmitter is off the air. One disadvantage is that the system must be taken off the air for several seconds to operate the coax switches.

A second method provided by some transmitter suppliers uses a pair of 3 dB hybrids interconnected with one fixed and one variable RF phasing section. The phasing section is constructed to operate while under RF power and can redirect the full output of either transmitter directly to the antenna and place the other transmitter into the test load without taking the system off the air. A dedicated system controller allows automatic or manual control. This so-called *switchless* combiner offers the highest possible on-air availability for combined FM transmitters. With complete redundancy in the RF power amplifier chain, some stations go one step further and also install dual exciters with automatic switching so that, if one exciter fails, the other unit is quickly switched into service.

90° Hybrid Couplers

Hybrid couplers are reciprocal four-port devices used either for splitting or combining RF sources over a wide frequency range. Figure 20 shows an exploded view of a typical 3 dB, 90°, hybrid coupler. The coupler consists of two identical parallel transmission lines coupled over a distance of approximately one-quarter wavelength and are enclosed within a single outer conductor. Ports at the same end of the coupler are in phase whereas ports at opposite ends of the coupler are in quadrature (90° phase shift) with respect to each other.

The phase shift between the two inputs or outputs is always 90° and is almost independent of frequency. If the coupler is being used to combine two signals into one output, these two signals must be fed to the hybrid coupler in phase quadrature. The reason this type of coupler is also called a 3

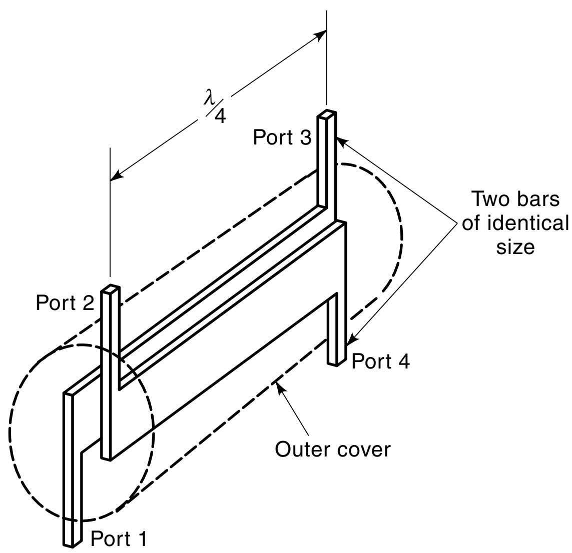

Figure 20. Physical model of a 90° hybrid coupler.

dB coupler is that when used as a power splitter, the split is equal or half-power (3 dB) between the two outputs.

90° Hybrid Combiners

The output hybrid combiner effectively isolates the two amplifiers from each other. Tuning adjustments can be made on one amplifier including turning it on and off without appreciably affecting the operation of the other amplifier. Good isolation is necessary so that if one transmitter fails, the other continues to operate normally instead of in a mistuned condition. Two of the ports on the hybrid coupler are the inputs from the power amplifiers, the sum port is the antenna output terminal, and the difference port goes to a resistive dummy load called the *reject load* because only the rejected power caused by imbalance appears here. When the power fed to each of the two inputs is equal in amplitude with a phase difference of 90°, the total power is delivered to the sum port (antenna). Very little of the power appears at the reject load if the phase relationship and power balance are correct. If the phase relationship is reversed between the two amplifiers, all the power is delivered to the reject load, so care must be taken to ensure that the proper one of the two possible 90° phase relationships is used. When all the ports on the hybrid combiner are properly terminated, isolation of 30 dB or more can be achieved between the power amplifiers. For perfect isolation between the amplifiers, the load impedance on the sum and difference ports must be exactly the same. This is approached in practice by providing a 1.0 : 1 VSWR with a resistive 50 Ω load for the termination (reject load) on the difference port and then reducing the VSWR on the antenna transmission line as low as possible by trimming the antenna match. This keeps the input port impedances from changing very much when one amplifier is not operating.

The input ports will present a load to each transmitter with a VSWR that is lower than the VSWR on the output transmission line because part of the reflected power coming into the output port will be directed to the reject load and only a portion will be fed back into the transmitters. Figure 21 shows the effect of output port VSWR on the input port VSWR and on the isolation between ports.

If the two inputs from the separate amplifiers are not equal in amplitude or exactly in phase quadrature, some of the power will be dissipated in the difference port reject load. The match in input power and phase is not extremely critical as shown in Figs. 22 and 23. The power lost in the difference port reject load can be easily reduced to a negligible value by touching up the amplifier tuning and by adjusting the phase shift. For example, if one amplifier is delivering only half the power of the other amplifier, only about 3% of the total available power will be dissipated in the reject load and 97% is still fed to the output transmission line (17).

If one transmitter fails completely, half of the working amplifier's output goes to the antenna, and the other half is dissipated in the difference port reject load. This is why the radiated output drops by 6 dB or to one-fourth of the original

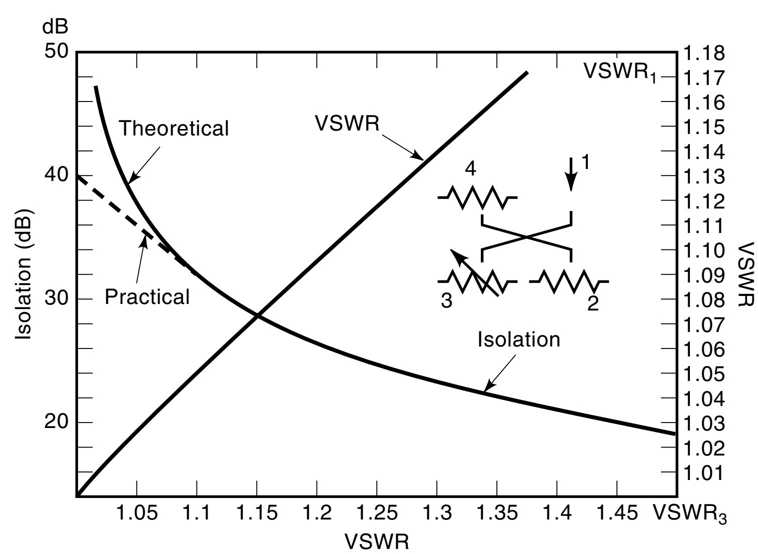

Figure 21. Isolation and VSWR of a 90° hybrid coupler.

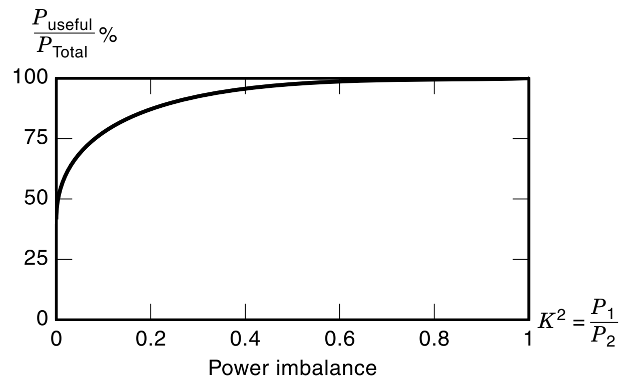

Figure 22. Loss due to power imbalance in 90° hybrid coupler.

combined power. The reject load must be rated to handle a minimum of one-fourth of the total combined power, but often the reject load is rated to handle one-half the total power, so that it can also be used a test load for one of the transmitters.

Hybrid Splitting of Exciter Power

Figure 24 shows a block diagram of a pair of combined amplifiers with dual exciters. The exciters cannot be operated in parallel like the amplifiers because their RF outputs would have to be on exactly the same carrier frequency and exactly in phase under all modulation conditions. An automatic or manual exciter switcher is used to direct the output of the desired exciter to the combined transmitter, and the other stand-by exciter is routed to a dummy load. The one exciter in use feeds a hybrid splitter/phase shifter which transforms one 50 Ω input into two isolated 50 Ω outputs that have a 90° phase shift between them with half the power going to each output. The operation of this hybrid splitter is the reciprocal of the hybrid combiner described above. The exciter must have enough power output capability to drive both power amplifiers. In some cases an additional IPA is required between the exciter and the splitter to boost the drive level. The length of coax from the power splitter to each amplifier input must be cut to a precise length so that the amplifiers will be fed in the proper phase relationship.

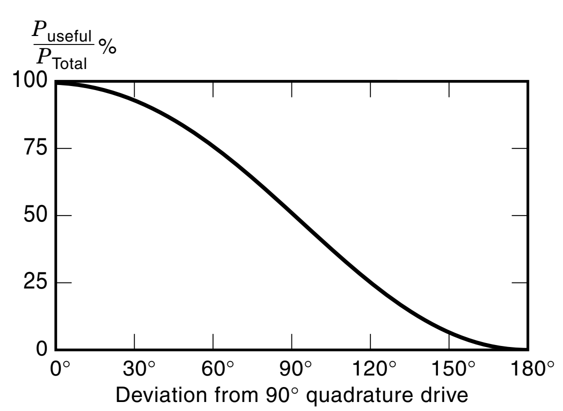

Figure 23. Phase sensitivity in a 90° hybrid coupler.

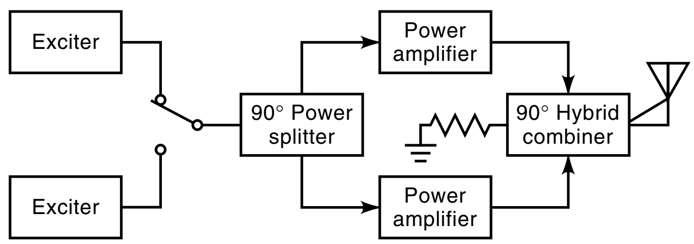

Figure 24. Block diagram of a transmitter with two power amplifiers, a 90° hybrid combiner, and dual exciters with 90° power splitter.

Each of the power amplifiers is assumed to have equal gain and phase shift. In practice, it may be difficult to tune the amplifiers so that their gains and phase shifts are equal at the same time. For this reason, a line stretcher or variable phase-shift network is usually included with the exciter splitter so that the station engineer can adjust phasing independent of amplifier tuning.

Filterplexing

The practice of having several FM stations share a single broadband antenna system has become more popular in recent years. To connect several transmitters on different frequencies together onto one antenna system, a special device called a filterplexer is required. The purpose of the filterplexer is to provide isolation between the various transmitters while efficiently combining their power into a single transmission line. This is usually accomplished by a system of band-pass filters, band-reject filters, and hybrid combiners. The isolation is required to prevent power from one transmitter from entering another transmitter with resulting spurious emissions and to keep the rest of the system running in the event of the failure of one or more transmitters.

An important consideration in designing a filterplexing system is the effect on the phase response (group delay characteristic in the passband) of each of the signals passing through the system because of individual bandwidth limitations on each of the inputs.

RF Intermodulation Between FM Broadcast Transmitters

Interference with other stations within the FM broadcast band and with other services outside the broadcast band can be caused by RF intermodulation between two or more FM broadcast transmitters. Transmitter manufacturers have begun to characterize the susceptibility of their equipment to RF intermodulation so this information will be available to the designers of filterplexing equipment.

The degree of intermodulation interference generated within a given system can be accurately predicted before the system is built if the actual mixing loss of the transmitters is available when the system is designed. Accurate data on *mixing loss* or *turnaround-loss* speeds the design of filterplexing equipment and also results in higher performance and more cost effective designs because the exact degree of isolation required is known before the system is designed. Filterplexer characteristics and antenna isolation requirements can be tailored to the specific requirements of the transmitters being used. The end user is assured in advance of construction that

Third-order intermodulation products

$f_1 = 100.3$ MHz. $f_2 = 101.1$ MHz.

$$2f_1 - f_2 = [2(100.3) - (101.1)] = [200.6 - 101.1] = 99.5 \text{ MHz.}$$

$$2f_2 - f_1 = [2(101.1) - (100.3)] = [202.2 - 100.3] = 101.9 \text{ MHz.}$$

OR

$$[f_1 - (f_2 - f_1)] = [100.3 - (101.1 - 100.3)] = [100.3 - 0.8] = 99.5 \text{ MHz.}$$

$$[f_2 + (f_2 - f_1)] = [101.1 + (101.1 - 100.3)] = [101.1 + 0.8] = 101.9 \text{ MHz.}$$

Figure 25. Calculation of intermodulation product frequencies.

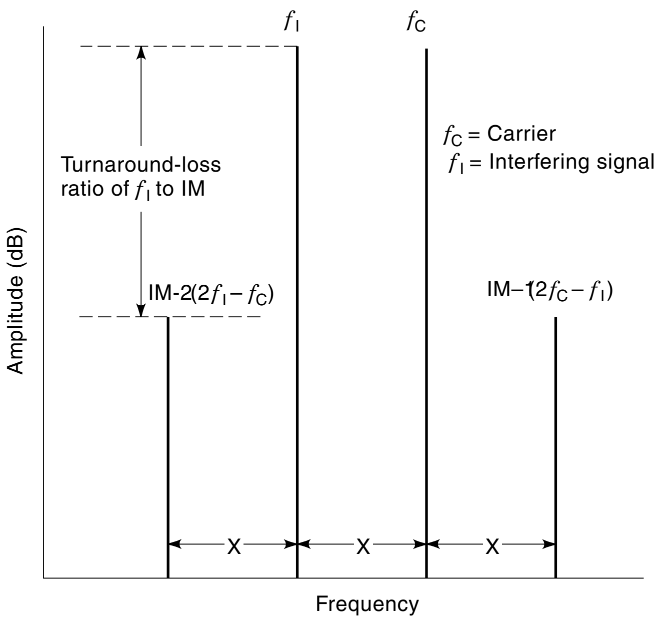

Figure 26. Frequency spectrum of a third-order IM products with the interfering level equal to the carrier level.

the system will perform to specification without fear of overdesign or underdesign of the components within the system.

Mechanisms Which Generate RF Intermodulation Products

When two or more transmitters are coupled to each other, new spectral components are produced by mixing the fundamental and harmonic terms of each of the desired output frequencies. For example, if only two transmitters are involved, the third-order intermodulation (IM3) terms could be generated in the following way. The output of the first transmitter f_1 is coupled into the nonlinear output stage of the second transmitter f_2 because there is not complete isolation between the two output stages. f_1 will mix with the second harmonic of f_2 producing an in-band third-order term with a frequency of $[2(f_2) - (f_1)]$. Similarly, the other third-order term will be produced at a frequency of $[2(f_1) - (f_2)]$. This implies that the second harmonic content within each transmitter's output stage along with the specific nonlinear characteristics of the output stage will have an effect on the value of the mixing loss.

It is possible, however, to generate these same third-order terms in another way. If the difference frequency between the two transmitters $[(f_2) - (f_1)]$, which is an out-of-band frequency, remixes with either (f_1) or (f_2), the same third-order intermodulation frequencies are produced.

Empirical measurements indicate that the $[2(f_2) - (f_1)]$ type of mechanism is the dominant mode generating third-

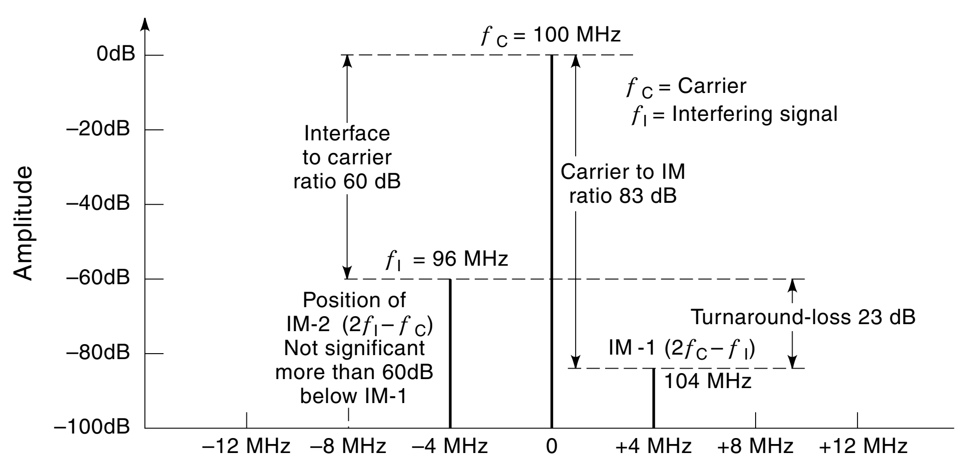

Figure 27. Typical frequency spectrum of a third-order IM products of a broadcast FM transmitter.

order IM products in modern transmitters using a tuned cavity for the output network.

Figure 25 is an example of calculating intermodulation product frequencies. Figures 26 and 27 show the resulting frequency spectra.

Intermodulation As A Function of *Turnaround Loss*

Turnaround loss or *mixing loss* describes the phenomenon whereby the interfering signal mixes with the fundamental and its harmonics within the nonlinear output device. This mixing occurs with a net conversion loss, hence the term *turnaround loss* has become widely used to quantify the ratio of the interfering level to the resulting IM3 level. A *turnaround loss* of 10 dB means that the IM3 product fed back to the antenna system will be 10 dB below the interfering signal fed into the transmitter's output stage.

Turnaround loss increases if the interfering signal falls outside the passband of the transmitter's output circuit, varying with the frequency separation of the desired signal and the interfering signal because the interfering signal is first attenuated by the selectivity going into the nonlinear device and then the IM3 product is further attenuated as it comes back out through the frequency selective circuit.

Turnaround loss can be broken down into three individual parts:

1. The basic in-band conversion loss of the nonlinear device;
2. The attenuation of the out-of-band interfering signal caused by the selectivity of the output stage; and
3. The attenuation of the resulting out-of-band IM3 products caused by the selectivity of the output stage.

As the *turnaround loss* increases, the level of undesirable intermodulation products is reduced, and the amount of isolation required between transmitters is also reduced.

The transmitter output circuit loading control directly affects the power amplifier source impedance and therefore af-

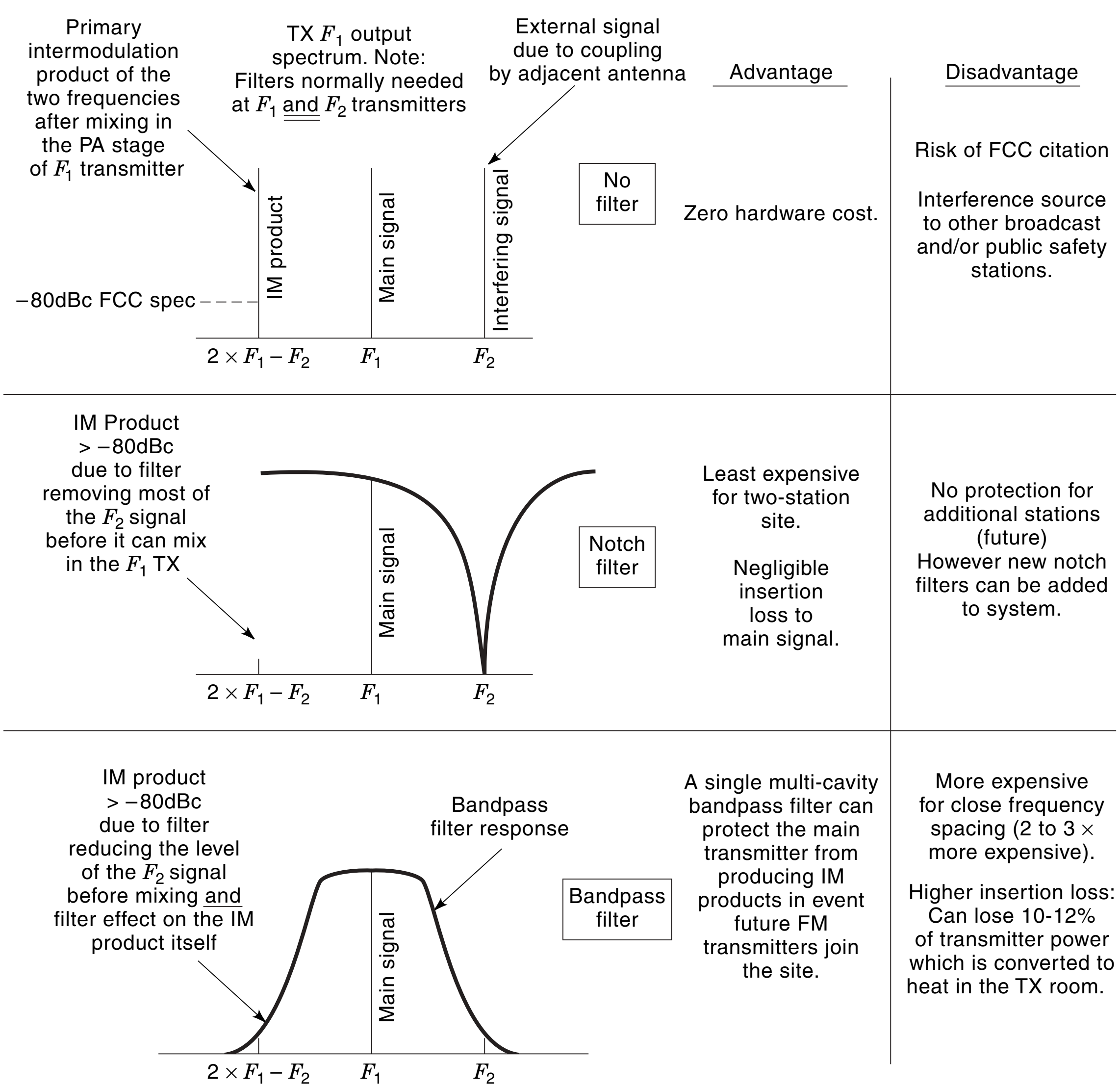

Figure 28. An overview of the various filtering options for preventing excessive IM3 products.

fects the efficiency of coupling the interfering signal into the output circuit where it mixes with the other frequencies present to produce IM3 products. Light loading reduces the amount of interference that enters the output circuit with a resulting increase in *turnaround loss.* In addition, the output loading control setting will change the output circuit bandwidth (loaded Q) and therefore also affect the amount of attenuation that out-of-band signals will encounter passing *into* and *out of* the output circuit (18).

Second harmonic traps or low-pass filters in the transmission line of either transmitter have little effect on the generation of intermodulation products because the harmonic content of the interfering signal entering the output circuit of the transmitter has much less effect on IM3 generation than the harmonic content within the nonlinear device itself. The resulting IM3 products fall within the passband of the low-pass filters and outside the reject band of the second harmonic traps. So these devices offer no attenuation to RF intermodulation products.

Figure 28 gives an overview of the various filtering options for preventing excessive IM3 products.

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