Active and Passive Components for Broadband Transmit Phased Arrays

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Phased-array antennas have penetrated a large application space, and in this article, we present some of the critical components for phased-array front ends, as described in Figure 1. We focus on an octave-bandwidth 6- to 12-GHz transmit implementation, but the methodology and component designs are scalable in frequency. Over the past few decades, phased-array applications have expanded from radar to wireless and satellite communications, as well as medical imaging and various sensing modalities. By tracking users with individual high-gain beams, spatial multiplexing...
enables cochannel interference mitigation, enhancing signal levels, and power management in wireless communications systems. For example, [1] shows a 64-element phased-array transceiver operating in the 28-GHz band with 52-dBm effective isotropic radiated power (EIRP) for 5G applications with up to 12 Gb/s of throughput, and [2] demonstrates a system-level development of a 384-element, 16-tile, W-band phased array with 52-dBm EIRP in 0.18-μm silicon–germanium bipolar CMOS (BiCMOS) technology. In satellite broadcasting and communication systems, phased-array antennas enable beam shaping over countries and continents, tracking at the end-user side, and rain fading mitigation, e.g., [3]. In [4], advances in phased arrays for satellite applications developed at Lockheed Martin are described from an industry perspective.

Phased arrays also provide enhanced capabilities for airborne radar: power efficiency increase; adaptive beamforming to suppress jammers; space-time adaptive processing to suppress clutter return; multibeam on receive capability for multistatic operation in stealth target detection; and higher system availability due to graceful degradation of array performance resulting in low lifecycle cost [5]. For fast mobile applications, automotive radars are used as sensors for adaptive cruise-control (ACC), safety applications like blind-spot surveillance and collision mitigation/avoidance, backing/parking, and low-speed/stop-and-go ACC [6]. In [7], a system simulation using a phased-array frequency-modulated continuous-wave radar with beam steering controlled by environmental variables is implemented, demonstrating the advantages of increased range and providing more time for adaptation. The automotive environment presents design challenges such as packaging and a harsh thermomechanical environment. In [6], automotive radar front-end phased array are analyzed, where RF microelectromechanical systems (MEMS) are used as switches in a Rotman lens or in phase shifters and optimized for RF and thermomechanical behavior.

For weather monitoring, phased-array radars bring the capability of scanning bigger volumes in shorter time intervals achieving update rates of <1 mm [8], where beam control and null generation enable collecting robust data for post-statistical analysis [9]. Also, phased arrays have found applications in RFID for localization in the ultrahigh frequency range, e.g., [10], while in radio astronomy [11], adaptive nulling techniques help combat radio frequency interference [12]. More recently, beam-steering techniques have been extended to biomedical devices [13]. Doppler-based noncontact vital-sign (NCVS) sensor systems are used to monitor heart and respiration rates of patients. The sensor accuracy can deteriorate quickly in a noisy or cluttered environment, and [14] reports on directive phased arrays for NCVS systems with considerably improved accuracy compared to fixed-beam Doppler radar. In [15], a 70-MHz near-field phased array provides 3D specific absorption rate control and increases locoregional hyperthermia in tumor treatments, while in magnetic resonance imaging systems signal-to-noise ratio, image contrast and resolution can be improved [16]. Phased arrays are also applied in wireless power transfer systems for maximizing power delivery to an intended location while minimizing leakage fields in other locations, e.g., [17].

The growing variety of applications and system requirements lead to the need of scalable, modularized devices for transmit and receive chains. In this context, [18], e.g., shows heterogeneously-integrated phased arrays for communication and sensor applications at Ku-band. In the transmit chain, spatial power combining is used to increase the EIRP, where each element or subarray is fed by an independent RF front-end circuit. This approach allows operation with multiple signals and beams using the same fixed antenna aperture, with tradeoffs in size and performance requirements for high-power, wideband operation [19]. Although it is an effective method to increase radiated power, size and generated heat remain key limitation as frequency increases and the λ/2 spacing of the antennas shrinks smaller than the size of active elements. Therefore, power density and efficiency of the power amplifiers (PAs) favor the use of gallium nitride (GaN) over gallium arsenide (GaAs). The high breakdown voltage of GaN results in high peak power levels and is accompanied by a robust silicon carbide (SiC) substrate with good thermal conductivity. Trends in active arrays for space and radar applications, with GaAs replaced by GaN PAs to increase EIRP with the same number of elements are discussed in [20] and [21].

Instantaneous bandwidth and complex modulation schemes place broadband operation demands on microwave circuits. Although most phased arrays are relatively narrowband, a number of arrays with broadband optical beamforming, e.g., [22], and CMOS true-time delay beamformers, e.g., [23], have been demonstrated. A comprehensive overview of digital and hybrid approaches can be found in, e.g., [24]–[26].

This article addresses microwave and millimeter-wave active and passive components for broadband-phased array front ends, focusing mostly on the transmit path. Figure 1 shows a high-level block diagram that outlines the article content. The main blocks include the beamforming network, amplification stage, isolation stage between transmit and receive, and the final radiating aperture. The digital control and dc supply are not discussed here but are an integral part of the system. The last component in the chain is the passive antenna itself, and its radiating and matching
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Characteristics affect the entire system performance. More specifically, the coupling between antenna elements results in an active scan impedance that varies across the array and across frequency and scan angle, as overviewed in, e.g., [27].

We present previous and ongoing studies of the transmit-chain front-end components shown in Figure 1, including design approaches and performance tradeoff analyses. The “Beamforming Networks” section overviews three broadband analog beamforming approaches: octave-bandwidth monolithic microwave integrated circuit (MMIC) phase shifters based on loaded transmission lines; an octave-bandwidth Butler matrix; and a millimeter-wave frequency-steered array.

The “Isolation Circuits” section presents a 7–13-GHz MMIC active circulator and a broadband analog method for canceling self-interference in the receive path. In the “Analysis of PAs in a Phased-Array Environment” section, a broadband nonlinear analysis for the transmit chain is shown, including the effects of antenna element mismatch due to scanning. The “Discussion and Outlook” section introduces a quantum sensor with potential for broadband traceable electric field measurements as a method for near-field array calibration and a supply modulation as an efficiency enhancement technique for PAs on the example of a three-stage 18–25-GHz MMIC PA.

Throughout the article, we focus on III-V semiconductor (GaAs and GaN) MMICs for the active components with potential for heterogeneous integration.

**Beamforming Networks**

In the case of broadband-phased arrays, beam squint can be mitigated using true-time delay beamforming, which can be done in the optical or microwave domains. Common microwave true-time delay phase

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**Figure 1.** A high-level block diagram of a phased array, where the beamforming can be done either in the digital or analog domains. The dotted arrows indicate coupling between the antenna elements. The focus of this article is on analog components of broadband transmit arrays, including beamforming networks, amplification, isolation, and field measurements. Shown in the expanded blocks with photographs are MMIC phase shifters, a Butler matrix, and frequency scanning as the beamformer networks; MMIC PAs; MMIC active circulators; and field sensors for aperture electrometry. Past and ongoing research at the University of Colorado, related to the design of these components, is described throughout the article.
shifters include loaded lines [28] and switched delay lines [29]. In the next section, “Broadband GaAs and GaN MMIC Phase Shifters for Beamforming,” continuous analog phase shifters and a corporate feed network in the 6–12 GHz octave band are discussed. Discrete beams can be achieved with other true-time delay feed networks, such as Butler and Blass matrices, and Rotman and Luneburg lenses [27]. Later in this section, a Butler matrix covering the 6–12 GHz octave band is presented. These approaches do not scale favorably to higher millimeter-wave frequencies, where phase shifters and transmission-line elements are too lossy. Therefore, an approach that can use a lower-frequency VCO and frequency multiplier as the input is studied on the example of a 130–180 GHz microfabricated broadband sequentially fed frequency-steered array for planetary-landing radar, described in the last part of this section.

**Broadband GaAs and GaN MMIC Phase Shifters for Beamforming**

Loaded transmission lines can provide a true-time delay over as much as a decade bandwidth, where the phase is varied using varactors [29], MEMS [28] or ferroelectrics [30] and are typically limited by return loss variation over frequency. Reflective phase shifters with different hybrids are presented in [31], [32], and the variable reflective loads can be implemented with MEMS or varactors. Large continuous phase shifts in [33], [34], are accomplished with two cascaded phase shifters. To achieve a good match and large phase shift across an octave band (6–12 GHz), we demonstrate the circuit in Figure 2, which combines a reflective topology with a loaded-line variable reactance. To showcase the architecture and investigate nonlinear performance, a GaAs MMIC with a phase that increases over the band, and a GaN MMIC with maximum phase shift in the center of the band are designed [35].

The phase shifter MMICs are implemented in WIN Semiconductors’ GaAs enhancement (E-) mode PIH10 and GaN depletion (D-) mode NP15-00 pseudomorphic high electron-mobility transistor (pHEMT) processes. The periphery of the transistors is chosen to approximately scale with power density of the processes: 4 × 75 μm in GaAs (∼ 1.5 W/mm) and 2 × 75 μm in GaN (∼ 3.5 W/mm). Each variable capacitor in Figure 2 is implemented with a transistor with grounded source and drain terminals. For the GaAs E-mode devices, the capacitance varies with gate bias (−0.2 to 0.6 V) from 0.241 to 0.470 pF at 9 GHz. The GaAs phase shifter uses

**Figure 2.** (a) A topology of reflective loaded-line continuous broadband phase shifter. The coupled and through ports of the Lange coupler are variable reactances, as in a reflective phase shifter, implemented with tunable artificial transmission lines that use diode-connected transistors as variable capacitors. (b) 2.5 mm × 2.5 mm GaAs and 1 mm × 2.5 mm GaN MMIC phase shifter. Both reactive loads are implemented with lumped inductors and a shorted termination in the GaAs MMIC (in blue) and with transmission-line sections and an open termination in the GaN MMIC (in red). The gate directions for the chips are indicated with red arrows. The effects of bends in the Lange couplers are compensated with electrical length and bridge placement.
Lange couplers and three lumped 0.4-nH inductors and four transistors in each shorted loaded line. For the D-mode GaN devices at 9 GHz, with a gate bias range of –3 to –1 V, the capacitance variation is 0.163–0.281 pF. The GaN phase shifter loaded lines are open-circuited and use 14 transistors with electrically short (1.7° at 12 GHz) high-impedance (95 Ω) lines as series inductors.

Figure 3(a) shows measured phase as a function of bias for the two MMICs in small-signal operation. Over this range, both phase shifters are matched with a return loss better than 10 dB across the frequency range, as a result of the reflective circuit architecture. The insertion loss is between 3 and 7 dB across the frequency range, based on tuning level. The MMICs are also tested for increased input power levels (5–25 dBm),

Figure 3. (a) A measured and simulated phase as a function of gate bias for several frequencies. The black lines indicate lower (dotted) and upper (solid) biases for a useful voltage range. The results for the GaAs and GaN MMICs are shown on the left (in blue) and right (in red), respectively. Simulations are shown in dashed and measurements in solid line. (b) Harmonic generation as a function of input power for the GaAs (in blue) and GaN (in red) phase shifters at 9 GHz. At the low-power end, the measurements are limited by the sensitivity in the experimental setup.
to evaluate the power handling and nonlinearities. A frequency-dependent increase in loss and a reduction in phase change is observed for both MMICs, but at different power levels, as expected. The GaAs MMIC, above 10 dBm of input power, provides 50°–60° of phase shift but with increasing loss. The GaN MMIC has 20°–75° of phase shift for up to 20 dBm of input power with little variation in $|s_{21}|$ compared to the small-signal case. The power dependence eventually levels because the first diode behaves as a short and the phase is no longer tunable.

Figure 3(b) summarizes second and third harmonic generation behavior, showing that the GaN device is linear at 20-dB higher input power as compared to the GaAs device. This is expected and has benefits when higher drive power is needed for the final amplifier chain and nonlinearities in the phase shifter can affect the overall linearity.

In a phased array, the broadband phase shifters require integration in a broadband corporate feed network. A single-stage two-way Wilkinson power divider with a voltage standing-wave ratio (VSWR) of 2:1 has a 3:1 bandwidth. To increase the bandwidth further, multisection Wilkinson dividers can be implemented [36]. Air-based TEM waveguides offer lower loss and dispersion when compared to planar transmission lines such as microstrip [37]. Figure 4 shows a five-section greater than decade (11:1) bandwidth Wilkinson divider implemented in air-filled coaxial lines fabricated using the PolyStrata process, which involves sequential deposition of copper layers and photoresist on a silicon wafer [38]. The copper layer thicknesses range from 10 to 100 μm with aspect ratios larger than 1:5. After the desired layers are deposited, the photoresist is released leaving an air-filled low-loss structure, enabling low-loss air-filled coaxial lines up to W-band and waveguides at higher frequencies. The Wilkinson divider is measured with thru-reflect-line microcoaxial calibration and shows better than −13 dB match with a loss of 0.4–1.2 dB from 2 to 22 GHz [39]. Integration of a MMIC in the microcoaxial PolyStrata environment is demonstrated in [40], showing that MMIC phase shifters can be integrated in such a low-loss broadband beamforming network.

**Broadband Butler Matrix**

A Butler matrix, first reported in [41], is a network implemented with hybrid couplers and fixed phase shifters, as shown in Figure 5(a). The network has $2^N$ inputs and $2^N$ outputs, where $N$ is the size of the matrix. Depending on which input is excited, a different phase progression is seen at the output antenna ports, corresponding to beams in different directions. For a $4 \times 4$ Butler matrix shown in the figure, two 90° coupler pairs (one at the input and one at the output) with 45° phase shifts in two of the branches and two crossovers, produce an output phase progression of $\pm 45°$, $\pm 135°$. A broadband Butler matrix requires a low-reflection, low-loss, and high-isolation four-port crossover across a large bandwidth, not readily available in standard technologies. The coupler amplitude and phase balance is critical since these errors are cascaded through the two coupler paths, and a fixed phased shifter across a wide bandwidth also requires careful design.
Figure 5. (a) A schematic of a $4 \times 4$ Butler matrix with $90^\circ$ hybrids. Under the matrix is an implementation in the HRL MECA heterogeneous integration process with alumina passives. Note that another crossover is necessary at the output for connection to the antenna array. The insert shows a closeup of the crossover, which utilizes the additional dimension given by the interconnect layer in the MECA process. (b) A full-wave simulated phase difference to the output ports when Port 1 is excited and all other ports terminated, with Port 5 serving as a reference for the phase, showing approximately $135^\circ$ of phase progression. Under this is a far-field normalized H-plane radiation patterns of the Butler matrix feeding a broadband ridged horn antenna array when Port 1 (shown on the left) and Port 2 (shown on the right) are excited. Patterns are shown at the edge and middle of the band at 6 GHz (in red), 9 GHz (in blue), and 12 GHz (in black).
Broadband 4 × 4 Butler matrices in the approximate frequency range of interest are summarized in Table 1 and include a 3- to 10-GHz multilayer stripline in [42] where the ports are not arranged for direct connection to an antenna array, the phase varies ±14° and insertion loss degrades severely at higher frequencies. In [43], a corrugated slot-coupled structure was used for the broadband components from 3.1 to 10.6 GHz, with a resulting magnitude and phase variation of ±1.2 dB and ±4.5°. A 1.8- to 12-GHz stripline design using multisection couplers is shown in [44] with the magnitude and phase variation between the ports of ±2 dB and phase ±15°. More narrowband versions with good performance are shown in SIW, e.g., from 8.5 to 10.6 GHz [45]. A larger 8 × 8 matrix from 2 to 18 GHz is shown in [48], with increased loss and amplitude/phase variation, especially at higher frequencies.

The design in Figure 5(a) is implemented in HRL’s Laboratories metal embedded chip assembly (MECA) process [49], with broadband coupler and interconnects demonstrated in [50]. The goal of the design shown here is to minimize phase and amplitude imbalance over 6–12 GHz. The Lange couplers are folded to reduce footprint, the crossover and Schiffman fixed phase shifters are designed in the MECA interconnect layer, and additional delay lines provide the proper 45° difference between the middle and outer paths. This Butler matrix is designed in three separate pieces of alumina connected by broadband interconnects of the MECA process. Full-wave simulations are shown in Figure 5 for the phase progression referenced to Port 5 across the 6- to 12-GHz bandwidth with less than ±21° variation. In this range, the input ports are matched to better than 7 dB and the insertion loss is within 1.5 dB. Return loss is better than 10 dB above 7 GHz. This design would need another crossover between the inner two ports for connection to a 4 × 1 antenna array. Simulations with the antenna array made of broadband ridged horn antenna elements [51] result in radiation patterns shown in Figure 5(b). With Port 1 excited, which has phase progression of approximately 135°, at 6 GHz the beam steers 40°, while at 9 and 12 GHz the steering is 26° and 21°, respectively. With the phase progression corresponding to Port 2 of approximately 45°, at 6 GHz the beam steers by 14°, while at 9 GHz and 12 GHz a steering of 9° and 7° is obtained. This beam former is currently being fabricated and will be connected to the antenna array described in the “Analysis of PAs in a Phased Array Environment” section.

**TABLE 1. Summary of broadband Butler matrices.**

<table>
<thead>
<tr>
<th>Reference</th>
<th>f₁ (GHz)</th>
<th>f₈ (GHz)</th>
<th>Technology</th>
<th>Size</th>
<th>ΔΘ (°)</th>
<th>ΔA (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[42]</td>
<td>3</td>
<td>10</td>
<td>Stripline</td>
<td>4 × 4</td>
<td>14</td>
<td>Not given</td>
</tr>
<tr>
<td>[43]</td>
<td>3.1</td>
<td>10.6</td>
<td>Corrugated slot</td>
<td>4 × 4</td>
<td>4.5</td>
<td>1.2</td>
</tr>
<tr>
<td>[44]</td>
<td>1.8</td>
<td>12</td>
<td>Stripline</td>
<td>4 × 4</td>
<td>15</td>
<td>2</td>
</tr>
<tr>
<td>[45]</td>
<td>8.5</td>
<td>10.6</td>
<td>SIW</td>
<td>4 × 4</td>
<td>5</td>
<td>0.6</td>
</tr>
<tr>
<td>[46]</td>
<td>2</td>
<td>3</td>
<td>Stripline</td>
<td>4 × 4</td>
<td>3</td>
<td>Not given</td>
</tr>
<tr>
<td>[47]</td>
<td>8</td>
<td>12</td>
<td>Stripline</td>
<td>8 × 8</td>
<td>15</td>
<td>1</td>
</tr>
<tr>
<td>[48]</td>
<td>2</td>
<td>18</td>
<td>Stripline</td>
<td>8 × 8</td>
<td>15</td>
<td>1.5</td>
</tr>
</tbody>
</table>

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**The phase of each element is determined by the length of the feed line section between the elements as well as the mutual coupling.**

**Broadband Frequency Scanning**

For space applications such as planetary-landing radar [52], higher millimeter-wave frequencies offer reduced mass but may not suffer attenuation as in the Earth’s atmosphere. At W-band and above, MMIC phase shifters, e.g., [53], and Butler matrices, e.g., [54], become very lossy and require specialized technologies. Another approach is frequency steering, which requires a voltage-controlled oscillator and potentially also a frequency multiplier, possible in MMIC technologies. Frequency-scanned waveguide arrays have been used since the 1940s, e.g., [55], and require broadband components to achieve a large scan angle. Figure 6(a) shows a linear array where the elements are fed with a varying frequency, which changes the relative phase between elements resulting in beam steering. The input wave travels along the feed line and loses power at each element to radiation. Elements closer to the feed input couple weakly to the feed line, while elements near the load couple strongly and any remaining power after the last element is dissipated in the load (typically 10–15% of the input power). The phase of each element is determined by the length of the feed line section between the elements as well as the mutual coupling. In Figure 6(a) the array is fed by a line of physical length α between elements and propagation constant β, with element spacing of d in air. For an angle θᵢ of the m-th beam
peak, the phase is given by $\sin \theta = \frac{\lambda_0}{d} (\alpha / \lambda_g - m)$, where $\lambda_0$ and $\lambda_g$ are the free-space and feed guided wavelength, respectively. Both a longer feed line section and a more dispersive delay line result in a wider beam scan with frequency.

Two W-band and G-band architectures, based on microcoaxial line fed slots and slotted waveguides, are introduced in [56]–[59]. The devices are fabricated using the PolyStrata process with coaxial lines at W-band, or at higher frequencies using lower-loss rectangular waveguides, which are comparable in size to the microcoax, but have higher dispersion and lower loss. An example 20-element linear waveguide-fed slot array, which scans from $-40^\circ$ to $0^\circ$ when the frequency varies from 130 to 180 GHz is shown in Figure 6(b), along with measured radiation patterns. This corresponds to approximately $1^\circ$/GHz of scanning with gain and scan angle over frequency shown in Figure 6(c). With the same approach and by introducing corrugations in the bottom waveguide feed wall, the dispersion is increased to achieve $2^\circ$/GHz in a $32 \times 32$ 2D array with 1D scanning [59].

**Isolation Circuits**

Phased arrays are usually used to both transmit and receive, in which case some isolation method is required to protect the low-noise amplifier from saturating due to leakage from the transmitter. For nonsimultaneous transmit and receive applications, switches can have high isolation and transmitter

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**Figure 6.** (a) The principle of operation of frequency-scanned arrays where the phase at each radiating element depends on element spacing and dispersion of the feed line. (b) A microfabricated 20-element Polystrata WR-5 waveguide slotted array. (c) Measured normalized radiation patterns from 130 to 180 GHz and measured and simulated scan angle and gain over the frequency range [57].
amplifier bias can be turned off during receive mode. For simultaneous transmit and receive, circulators or diplexers are used, depending on the frequency of the up/down links. Often the required levels of isolation are greater than 100 dB and exceed what can be done by only one method, so several methods including e.g., antenna polarization are added, and have been demonstrated over a wide bandwidth, e.g., [60]. At the circuit level, circulators are bulky and usually narrowband, so a MMIC solution is attractive. For self-interference and other cancellation over a wide bandwidth, tunable notch filters are useful but are large and the additional loss degrades the noise figure. An active analog canceler amenable to monolithic integration can also produce a tunable spectral notch over a wide bandwidth. Circulator and analog cancellation GaAs MMICs that operate over an octave bandwidth are presented next.

**Active Circulator MMIC**

Nonreciprocal behavior of transistors can be used to achieve circulation and was first presented in 1965 [61]. Later, a variety of architectures were further explored, e.g., [62]. In [63], a MMIC implementation using Qorvo’s 0.25 µm GaAs pHEMT process, designed to operate in the 7–13 GHz band, is demonstrated (Figure 7). Circulation is accomplished by connecting three unconditionally stable, gain-matched, single-stage equal amplifiers with three asymmetric Lange couplers. The through ports are connected to the amplifier inputs, and the coupled ports to the outputs. The isolated ports are terminated in 50-Ω resistors and the Lange coupling coefficient specified to achieve a tradeoff between backward isolation ($|S_{31}|$) and forward gain ($|S_{21}|$), while ensuring stability. The forward gain of the circulator is proportional to the amplifier gain, reduced by the coupling and through attenuation of the

Figure 7. (a) A broadband 7- to 13-GHz active circulator topology. (b) A photograph of an active circulator GaAs MMIC circuit. When Port 1 (P1) is the input, P2 is the output, and P3 isolated. (c) Simulated and measured results for 50-Ω loads at all ports from 6 to 14 GHz showing insertion gain of up to 2.4 dB with isolation $|S_{31}|$ greater than 13 dB across the band [63].
Lange couplers. The simulated versus measured performance is shown in Figure 7(c). Simulations for 50-Ω loads at the ports shows insertion gain of up to 3 dB across 6–14 GHz with isolation \(|S_{31}|\) greater than 13 dB. Measurements show good comparison to simulated insertion gain, but a decrease in isolation, explained later in this article.

The 2.5 mm × 2.5 mm chip [Figure 7(b)] behavior is analyzed when integrated in a phased array–like environment, where the port impedances vary and where process variations impact device performance [Figure 8(a)]. The circuit architecture contains many possible feedback paths, so stability needs to be carefully analyzed. Assuming that all port impedances can vary within a VSWR = 2, Figure 8(b) shows simulated statistical analysis of the loop gain for the three amplifiers in the circulator. A safe phase margin is obtained from dc to 20 GHz, indicating stability over a wide range of loads. To explain the degradation in isolation measured in Figure 7(c), a statistical analysis is performed with information provided by Qorvo for the GaAs process variations, and with additional load mismatch on all three ports inside the VSWR = 1.1 circle with a normal distribution of impedances. The results in Figure 8(c) show a large spread in the three isolation parameters. For some values of process parameters and load impedances, the degradation in isolation seen in the measurement is predicted.

**Active Broadband Interference Cancellation**

Agile broadband receivers are susceptible to self or external interference where the analog to digital converter (ADC) can become overloaded. A possible solution is the use of electronically-tunable passive filters, extensively reported in the literature, e.g., [64]–[66], however they can be lossy and bulky. Another solution is a cancellation circuit amenable to MMIC integration and co-integration with other front-end RF circuits. One
example is analog finite impulse response filters, which have been implemented in CMOS and GaAs and used as equalizers in optical communication systems [67]–[69]. A similar idea has been used in RFID for leaking carrier suppression, where an interferometer circuit is designed for the carrier signal [70]. Here we extend these ideas to an active circuit that applies vector addition to eliminate unwanted signals, while relying on a single input signal.

The general interferometer circuit topology shown in Figure 9(a) creates tunable spectrum notches that can be placed over the largest interferers to reduce the dynamic range requirements on the ADC. The Lange couplers divide and recombine the input signal; one path has only a gain stage, while the other path also includes a variable lumped-element transmission line with variable shunt capacitances. In an MMIC, these are implemented with gate-source grounded transistors, which act as varactors. By changing the bias on the transistors, the delay through the artificial transmission line is varied to achieve destructive interference between the two paths at a given frequency. The MMIC shown in Figure 9(b) is designed in the WIN Semiconductors’ GaAs E-mode single-polarity supply PIH10 process to operate across X-band. The two gain stages are biased slightly different with more gate voltage applied to the path with the transmission line to account for the additional loss. Figure 9(c) shows measured behavior when the control voltage for the variable transmission line varies from −0.7 V to 0.7 V. The circuit is well matched from 6 GHz to 12 GHz, while the null can be moved from 8.2 GHz to 10.8 GHz. The measured behavior matches the simulated behavior, except for a 0.4 GHz downward shift in tuning range frequency.

Analysis of PAs in a Phased-Array Environment

In transmit phased array, e.g., for jamming applications, PAs are the last active component in the chain preceding the antenna elements. This section presents an analysis of such a chain, noting that the effects of a switch, circulator, or diplexer can be added without loss of generality. Figures-of-merit (FoM) for broadband transmit electronically scanned arrays are highly dependent on both passive and active component variation over frequency and power as well as their interactions. The FoMs we consider here are related to the circuit side of the array, and include impedance variation due to antenna coupling, power-added and overall efficiency, and gain/loss. The antenna-related FoMs are the EIRP and scan volume, not discussed here. We first consider how the coupling between elements affects the PA efficiency, on the example of a small horn antenna presented in [51] and shown in Figure 10(a) and (b) together with its measured and simulated octave-bandwidth performance. This antenna element is characterized, built into a 4 x 1 linear array [Figure 10(c)] and is a starting point for analyzing effects of impedance variation on the PAs that feed the four ports in the active array.

Figure 11(a) shows the block diagram used in simulation for coupling analysis as a function of the scan angle assuming ideal phase shifters. The full-wave electromagnetic (EM) simulations (HFSS) are imported into Cadence/AWR Microwave Office circuit simulations to determine the coupling as a function of progressive phase shift across the array. A coupler in each element is used to measure the complex backward propagating wave from the antenna through the phase shifter. The power of this wave in...
each port, when all other ports are fed with a nominal 0 dBm of input power, is shown in Figure 11(b) across frequency. As expected, the two inner and two outer elements show symmetric behavior. The ratio of the complex reflected wave and the incident wave gives the reflection coefficient, and therefore impedance presented to each amplifier when the other ports are active.

The PA used for the analysis is a 6- to 12-GHz Qorvo QPA2598 GaN MMIC die, with a peak output power of 2 W, small-signal gain of 23 dB and power-added efficiency (PAE) = 30% in continuous wave (CW) operation at a drain bias of 25 V. Qorvo provided an X-parameter model for this device, with simulated large-signal parameters shown in Figure 12. The coupling between the output ports of the PAs, associated with a progressive phase shift for different steering angles, forces active impedance trajectories shown in the Smith charts in Figure 13 for the edges and center of the octave band, and given by

$$\Gamma_{1,\text{active}} = \frac{b_1}{a_1} = S_{11} + S_{12} \frac{b_2}{a_1} + S_{13} \frac{b_3}{a_1} + S_{14} \frac{b_4}{a_1}.$$ 

The simulation setup is shown in Figure 13(a), and the active impedance coupling matrix as a function of progressive phase shift for $\phi = [-180^\circ, 180^\circ]$ computation [Figure 13(b)] is used to evaluate the PAE load-pull performance of the linear active array at 1 dB compression of each of the PAs. Note that the maximum PAE is obtained for different impedances across the bandwidth, ranging from close to 50 $\Omega$ at 6 GHz to an inductive impedance at 9 GHz and a lower real impedance at 12 GHz. The design attempted a flat PAE at 50 $\Omega$ and over frequency. In a 4 x 1 array, outer and inner elements have the

![Figure 10](image-url)

**Figure 10.** (a) An octave-bandwidth small-aperture ridge waveguide horn with dielectric lens showing the waveguide-to-microstrip transition [71], E-plane ridge, and dielectric loading that improves the match. (b) A measured and simulated return loss and gain from 6 to 12 GHz [51]. (c) The horn elements are combined in a 4 x 1 linear active array with broadband GaN MMIC PAs feeding ports P1–P4.
same Γ trajectory behavior, but occupy symmetric loci in each path, so the PAE and $P_{\text{out}}$ performance is different for each element [Figure 13(c)]. The PAE of the four amplifiers as a function of progressive phase shift $\phi = [-60^\circ, +60^\circ]$ is shown in Figure 14 across frequency.

**Discussion and Outlook**

Broadband phased array implementations present numerous challenges, and this article only touches upon some front-end components that offer performance over octave or larger bandwidth in the microwave and millimeter-wave ranges. There are a number of other useful advances in the areas of array calibration and high-efficiency PAs with dynamic supplies, briefly overviewed next.

**Sensors for Electric Field Measurements**

Calibration of phased arrays often involves measurements of near-field magnitude and phase across the aperture, e.g., [72] and [73]. Recent interest in quantum sensors for microwave electrometry have resulted in a number of demonstrations of both room-temperature [74]–[76] and cold [77] Rydberg-atom field measurements. These alkali atoms in highly excited states can be used to make traceable microwave electric field

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**Figure 11.** (a) Simulation setup where the full-wave EM simulated 4-element linear array is imported into a circuit simulation with ideal phase shifters, couplers, and switches to determine coupling between elements. (b) Simulated coupling over frequency and progressive phase shift for the four elements of the array at ports P1–P4 indicated in Figure 10(c).
measurements and require a calibration factor as shown in [74]. The atoms in a vapor cell are excited with two lasers ("coupling" and "probe") at specific wavelengths making them sensitive to specific microwave frequencies. In principle, with widely tunable laser wavelengths, a practically continuous bandwidth from low GHz to millimeter-waves can be sensed. Rydberg atoms potentially increase sensitivity of electric field measurements over traditional probes and detectors down to 780 pV/m [78]. The electric field is extracted from a frequency measurement traceable to the fundamental Plank constant \( h \) as \( |E| = (h\Delta f_m)/\rho \), where \( \rho \) is the dipole moment of the atomic transition, and \( \Delta f_m \) is the change in frequency of the spectral splitting (known as Autler-Townes splitting). A photograph of a room-temperature Rubidium Rydberg-atom vapor cell and electric field measurements derived from Autler-Towns splitting are shown in Figure 15. The cell is made of glass and can be made very small, while the spatial resolution of the electric field detector is determined by the size of

Figure 12. Output power, power gain, and PAE for the 2-W Qorvo QPA2598 GaN MMIC PA, simulated with an available X-parameter model obtained from measurements. The simulation conditions are at a drain voltage \( V_{dd} = 25 \text{V} \) and with 50-\( \Omega \) port terminations.

Figure 13. (a) Array geometry used for computing the coupling matrix in HFSS. (b) Simulation schematics for active impedance computation as a function of progressive phase shift for \( \phi = [-180^\circ, 180^\circ] \), and load-pull schematic for computing PA the PAE at 6, 9, and 12 GHz at 1 dB compression. (c) A Smith chart showing \( \Gamma_A \) trajectory of inner and outer elements overlapped with PAE shown in color.
two intersecting laser beams. Recently demonstrated heterodyne measurements, e.g., [75] improve sensitivity and show potential of very sensitive electric field amplitude and phase measurements over a broad bandwidth.

**Enhanced-Efficiency Broadband PAs**

The PA used for simulations in the previous section is a commercial general-purpose broadband GaN MMIC with limited efficiency. PAs with enhanced efficiency through dynamic power supplies (envelope tracking [79]) have been demonstrated at higher power levels, with higher efficiencies and at increased frequencies. An example PA designed for K-band satellite communications from 18 to 25 GHz is described in [80]. This three-stage GaN MMIC fabricated in the Qorvo 150-nm process and shown in Figure 16(a) shows over 40% efficiency in CW with 4 W of output power from 18 to 24 GHz. The drain supply lines for the three stages can be accessed separately and by changing them dynamically with the envelope of a modulated signal, the average efficiency is improved. This approach of efficiency enhancement can be extended to high envelope bandwidths and multiple simultaneous signals over a large bandwidth.

**Conclusions**

This article presents several MMIC components amenable to heterogeneous integration for broadband-phased array, investigated by former and current Women in Microwaves students at the University of Colorado. GaAs and GaN octave-bandwidth 6–12-GHz MMIC phase shifters with a circuit topology that combines the reflective and loaded-line architectures are demonstrated for analog beamformers as well as a 6–12-GHz Butler matrix and a frequency-scanned broadband 130–180-GHz array. An active 6–14-GHz MMIC circulator is presented with potential of integration in GaN for increased power handling. For reducing unwanted interference, a GaAs MMIC with integrated low-noise amplifier can provide a broadband tunable notch in the 6–12-GHz range. Finally, the transmit portion of a four-element active array is analyzed to understand effects of element coupling on PAE over an octave bandwidth.

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**Figure 14.** PAE as a function of progressive phase shift across the 4-element linear array, across frequency and for all four amplifiers (PA1–PA4) driven 1 dB into compression.

**Figure 15.** An example of a measured electric field of a 17.04 GHz plane wave, copolarized with the light derived from Autler-Towns splitting measurements. The inset shows a photograph of Rubidium Rydberg-atom vapor cell made of Pyrex glass, with laser light input from fibers to excite a specific atomic transition corresponding to 17.04 GHz.
Ongoing research at the University of Colorado includes Rydberg-atom electrometry aimed at traceable phased-array calibration as well as broadband MMIC PAs with supply modulation for efficiency enhancement for simultaneous signals transmitted on different beams.

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**References**


Broadband micro-coaxial Wilkinson dividers


